

For Reference

NOT TO BE TAKEN FROM THIS ROOM

Ex LIBRIS
UNIVERSITATIS
ALBERTAENSIS





Digitized by the Internet Archive
in 2024 with funding from
University of Alberta Library

https://archive.org/details/Schmidt1982_0

THE UNIVERSITY OF ALBERTA

RELEASE FORM

NAME OF AUTHOR Christoph Schmidt

TITLE OF THESIS

Microprocessor

Control of an

Induction Motor

DEGREE FOR WHICH THESIS WAS PRESENTED Master of Science

YEAR THIS DEGREE GRANTED 1982

Permission is hereby granted to THE UNIVERSITY OF ALBERTA LIBRARY to reproduce single copies of this thesis and to lend or sell such copies for private, scholarly or scientific research purposes only.

The author reserves other publication rights, and neither the thesis nor extensive extracts from it may be printed or otherwise reproduced without the author's written permission.

THE UNIVERSITY OF ALBERTA

Microprocessor
Control of an
Induction Motor

by



Christoph Schmidt

A THESIS

SUBMITTED TO THE FACULTY OF GRADUATE STUDIES AND RESEARCH
IN PARTIAL FULFILMENT OF THE REQUIREMENTS FOR THE DEGREE
OF Master of Science

Electrical Engineering

EDMONTON, ALBERTA

Spring 1982

THE UNIVERSITY OF ALBERTA
FACULTY OF GRADUATE STUDIES AND RESEARCH

The undersigned certify that they have read, and
recommend to the Faculty of Graduate Studies and Research,
for acceptance, a thesis entitled

Microprocessor

Control of an

Induction Motor

submitted by Christoph Schmidt in partial fulfilment of the
requirements for the degree of Master of Science.

Abstract

This thesis presents a speed controller for an induction motor using a microcomputer. The supply for the motor consists of a three phase, pulse-width modulated voltage, which is synthesized by the microcomputer gating the thyristors in a dc-side commutated inverter. The possible applications for a variable speed drive using an induction motor are pointed out. The use of a microcomputer as controller for the inverter and the advantages of its use are described.

An introduction to the fundamentals of variable speed ac-drives is given, including a description of operating an induction motor from a nonsinusoidal supply. Various schemes for generating such a supply, and several choices for control strategies, are stated. The control strategy selected for the induction motor is the constant slip control. The inverter used in the study, as well as the characteristic features of the motor and the control computer, are described.

A description is given of the software for generating the gate pulses for the inverter and control and experimental results are included from laboratory tests conducted on a small motor in the University of Alberta Power Electronics Laboratories.

Acknowledgements

The author gratefully acknowledges the help and encouragement provided by his supervisor, Professor Ken E. Bollinger. His good sense of humor created a pleasant atmosphere in the Power Lab and the excellent research facilities provided by this Lab made working easier.

The author would like to thank Mr. Albert Huizinga and Roy Menezes for their support during the research project.

Further acknowledgements are to the University of Alberta and the German Academic Exchange Service (DAAD) for awarding the author the Killam Exchange Scholarship which created the basis for a successful research.

List of Symbols

a	turns ratio, stator to rotor windings
A	control signal
E	dc-voltage
e_c	capacitor voltage
Δ	error signal at time nT
f_1	stator frequency
f_2	rotor frequency
$f_2(n)$	rotor frequency at time nT
I_L	load current
I_M	maximum current
I_1	stator current
I_2	rotor current
K_P	gain factor of proportional controller
K_i	gain factor of integrating controller
K_d	gain factor of derivative controller
L_1	stray inductance, stator side
L_2	stray inductance, rotor side
m	number of phases
n	actual rotor speed
n_s	synchronous rotor speed
p	number of pole pairs
P_{diss}	dissipated power
R_1	resistance of stator windings
R_2	resistance of rotor windings
R_c	core-loss resistance
s	slip

t_{00}	thyristor turn-off time
V_b	blocking voltage
V_1	stator voltage
V_2	rotor voltage
V_{ph}	phase voltage
W_{comm}	stored energy per commutation
X_1	stator leakage reactance
X_2	rotor leakage reactance
X_m	magnetizing reactance
$X_{11} = X_1 + X_m$	stator reactance
$X_{22} = X_2 + X_m$	rotor reactance

Table of Contents

Chapter	Page
1. Introduction	1
2. Fundamentals of the Static Inverter Drive	7
2.1 The Induction Motor with Nonsinusoidal Supply Voltage	7
2.2 Inverter Types	11
2.2.1 The Cycloconverter	12
2.2.2 The Current Source Inverter	13
2.2.3 The Voltage Source Inverter	15
2.2.4 Pulse Width Modulator	16
2.3 Control Methods for Induction Motors	18
3. Description of the Inverter	25
3.1 The DC-Side Commutated Inverter Operation	25
3.2 The Commutation Sequence	29
3.3 The Firing Circuit	35
3.4 The Protection circuit	36
3.5 Design Considerations for the Inverter Components	38
4. Description of the Induction Motor and Related Equipment	42
4.1 The Induction Motor	42
4.2 The Control Computer	45
5. Program Description	48
5.1 Main Routine	48
5.2 Routine "start"	51
5.3 Routine "convert"	56
5.4 Routine "speed"	62
6. Experimental Results	68
6.0.1 Measurements	72

7. Conclusion	78
8. Bibliography	80
Appendix	86
8.1 Semiconductor Data-Sheets	86
8.2 Control Program Source Code	91

List of Tables

Table	Page
4.2 Motor nameplate reading	43
4.3 Generator nameplate reading	43
4.4 Locked rotor test	43
4.5 No load test	44
4.6 Load tests	44
4.7 Stator resistance measurement	45
4.8 Derived motor parameters	45
5.2 Commands for the motor control	48

List of Figures

Figure	Page
2.1	Equivalent diagram for the induction motor7
2.2	Input impedance as a function of slip8
2.3	Current as a function of slip8
2.4	Circle diagram for varying frequency9
2.5	Torque-speed characteristic10
2.6	Cycloconverter12
2.7	Current Source Inverter13
2.8	Voltage Source Inverter15
2.9	Output signal of the pulse width modulator16
2.10	Torque speed characteristic for voltage variation18
2.11	Torque speed characteristics with slip control23
2.12	Microcomputer implementation of the controlled slip drive24
3.1	Schematic of the dc-side commutated inverter26
3.2	Phase voltages and current waveforms for the six step inverter27
3.3	The voltages and currents at commutation30
3.4	Equivalent circuit at commutation instant31
3.5	Thyristor with snubber circuit34
3.6	Optically coupled firing circuit35
3.7	Overcurrent protection circuit36
3.8	The microcomputer to inverter interface circuit41
4.1	Circuit arrangement for measuring induction motor parameters42
5.1	Flowchart for the main routine50

Figure	Page
5.2 Subroutine "start"	54
5.3 Output voltage versus frequency	56
5.4 Inverter output voltage	56
5.5 Flowchart for the routine "convert"	61
5.6 Schematic of PID controller	62
5.7 Flowchart for routine "compare"	65
5.8 Flowchart for the routines "interrupt" and "overcurrent"	66
5.9 Flowchart for the routine "checkcurrent"	66
5.10 Flowchart for the routine "speed"	67
6.1 Experimental configuration for testing the PWM-inverter	69
6.2 Phase current and line to line voltage at 60 Hz	73
6.3 Phase current and line to line voltage at 15 Hz	73
6.4 Commutation capacitor current and thyristor blocking voltage	74
6.5 Commutation capacitor current and dc-bus voltage	74
6.6 Commutation capacitor current and dc-bus voltage at 60 Hz	74
6.7 Commutation capacitor current and dc-bus voltage at 15 Hz	74
6.8 Firing pulses Th, and ThC+ThD at 60 Hz	75
6.9 Firing pulses Th, and ThC+ThD at 15 Hz	75
6.10 Stepresponse for steps in reference	76
6.11 Rotor speed vs. time with softstart	76
6.12 Stepresponse for steps in load current	77

Figure

Page

6.13 Stepresponse with proportional control77

1. Introduction

The induction motor is the most common electromagnetic power transducer. It is estimated that over 60 % of the total electrical power consumption in North America is used by this type of motor, because it combines excellent performance at rated load and, especially with a squirrel-cage rotor, ruggedness and simplicity.

The induction motor though has two major disadvantages:

- the enormous amount of overcurrent when starting. (5 to 6 times rated current, mostly reactive)
- it is difficult to vary the speed of a squirrel-cage induction motor but with wound-rotor type motors a limited degree of speed variation can be achieved with a corresponding loss in efficiency.

The induction motor, therefore, is most economical at its rated speed.

Industrial processes, however, require an increasing number of variable speed drives, and the rising cost of energy determines the need for efficiency. In the past, wherever a variable speed drive was to be applied, the only choice available was a dc-motor because of lower cost. Problems associated with this type of motor surface when high-power drives are required or when the motor is to be applied in adverse environments (for instance submersed in fluids or where explosion hazards occur). The low inertia of the squirrel-cage induction motor, and the lack of mechanical elements that have to transfer the current to the

rotor (brushes), make this motor particularly suitable for the mining industry. The induction motor is more reliable, more compact and requires less maintenance (negligible maintenance in some cases) than the dc-motor.

Because of the rising cost of energy and the steadily decreasing cost of solid-state circuitry, variable speed ac drives have now become competitive with dc-drives and further improvements can be expected. In the conventional induction motor drive with fixed frequency, speed control is effected by inserting additional resistors in the phases of the rotor. This reshapes the torque-slip characteristic of the motor in such a way that the point of maximum torque is shifted to higher slip values (including $s=1$) without changing the maximum torque value.

But speed control by inserting additional resistors only works if there is a constant loading torque. Once this torque is removed the rotor approaches synchronous speed regardless of the size of the resistor inserted in the rotor. These disadvantages can be overcome by feeding the motor through a static inverter, rather than connecting it to the mains directly. With this inverter-motor configuration and an appropriate control strategy, many advantages can be achieved. These advantages include :

- adjustable speed from standstill up to rated motor speed or even higher,
- losses only slightly higher than those at rated speed as the motor speed is decreasing. (The losses mainly are

produced in the commutating circuitry of the inverter)

- constant speed even at varying loading torques,
- motor operation near the peak torque over the total speed range,
- possibility of soft starts with no excess currents, and
- no increase of current over the rated magnitude when starting.

Many publications can be found concerning the speed control of induction motors by analog means. Since the relationship between slip and torque is nonlinear at lower speed, some measures have to be applied to compensate for this nonlinearity. It is difficult to realize these correcting measures in analog circuits but this can easily be done with a programmable controller, using a microcomputer. Since the cost of microcomputers is rapidly decreasing and since they are easy to program, their use in controlling an induction motor-inverter configuration is justified. In using a microcomputer, there is also the possibility of applying optimal control methods by multivariable state-feedback control. This type of control is difficult to implement using analog hardware. Another advantage is that control parameters, like gain factors of the controller, can be changed interactively and on-line.

Applications for the induction motor - static inverter combination would be:

1. Large Fans

Large fans often have to be designed for the worst case,

or maximum air flow, but do not have to be run at that rate for most of the time. So the common practice is to run the fan only temporarily and whenever a measured variable (such as heat or exhaust gas) rises above a certain level, to start the fan. Every start produces enormous stress on the motor, since the starting current usually is 6 times the rated current and therefore the number of starts determines the lifetime of the motor. In addition, the large current is at a poor power factor and thus produces a considerable voltage drop in the power line. The inverter motor combination can solve these problems since, by running the motor at varying speeds not only can the air flow be adjusted, but also energy is saved and the lifetime of the motor can be extended.

2. Centrifugal Pumps

Centrifugal pumps for pipelines have to carry a number of different refinery products with different density coefficients. As described in [1], the working point of the pump is determined by the resistance characteristic of the pipeline. The conventional way of changing the flow rate is by altering the resistance of the pipeline through a throttling device. This causes the pump to assume another working point at the same, or only slightly changed, speed as before, but the throttling causes increased losses, both in the pipeline and in the motor.

With a speed controlled motor, energy can be saved. Instead of closing the valves for throttling the flow of the pumped medium, motor speed and power consumption are reduced.

3. Multi-motor Equipment

Soft starting of multi-motor equipment, such as large conveyor belts, with a fair amount of mass to be accelerated, can be achieved without oscillations and excess torque.

4. Other Applications

Applications also include situations where variable speed is desired and a dc-motor cannot be used, for maintenance reasons, explosion hazards, or because the required power is too much for a dc-motor to commute. The applications can be in mining, traction motors for large electric trains, petro-chemical sites, refineries and so on.

But with all these possible applications, the economical side of the coin must also be considered, since the price for a controlled speed drive is much higher than that for a single induction motor. Sometimes the amount of energy saved by the controlled speed drive is not substantial, especially when the power demand of the load does not follow a linear, but rather a square- or cube-law. Taking into account the overall efficiency of the inverter-motor-combination which will not exceed 86 % , even with the largest units, the losses in the uncontrolled motor will have to be well over 14 % to

justify the controlled speed drive for energy conservation reasons only.

2. Fundamentals of the Static Inverter Drive

2.1 The Induction Motor with Nonsinusoidal Supply Voltage

The lumped parameter diagram of the induction motor as shown in Figure 2.1, is the same as for a transformer, except that R_2/s replaces the load resistance.

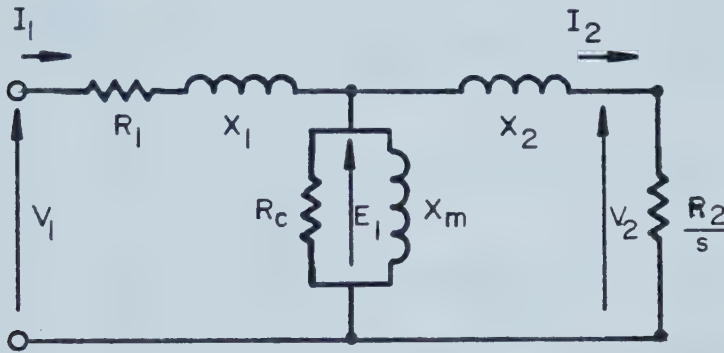


Figure 2.1 Equivalent diagram for the induction motor

As can be seen from the above diagram, the equation for the primary current can be expressed as:

$$I_1 = \frac{V_1}{R_1 + \frac{R_2}{s} + j(X_1 + X_2)} \quad (2.1)$$

This equation is readily derived from basic machine theory and includes the assumption that the magnetizing current is negligible. Equation 2.1 is used for deriving the circle diagram of the induction motor. From this diagram all characteristic data (current, power and torque) can be obtained by measuring the length of certain lines on the circle diagram and multiplying them by the appropriate scale

factor. The circle diagram shows the locus of the tip of the phasor I_1 and can be obtained by impedance inversion as described in reference [2].

The input impedance diagram for an induction motor is shown in Figure 2.2.

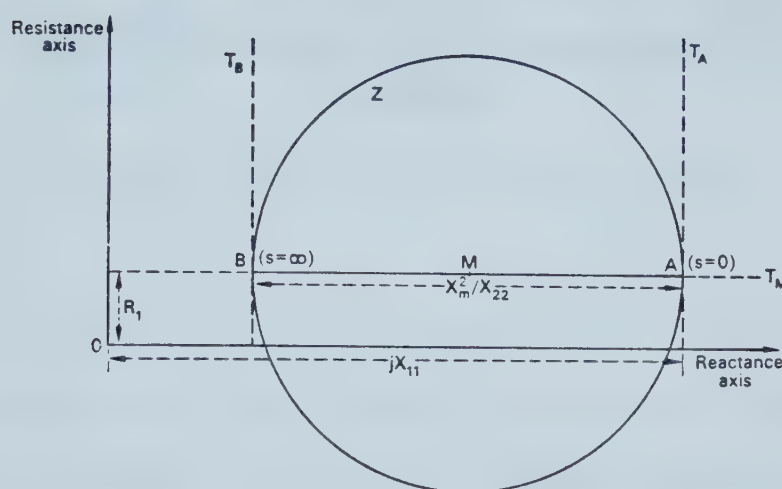


Figure 2.2 Input impedance as a function of slip [2].

The only parameter is s or R_2/s and a negative slip means that the rotor is driven externally into supersynchronism.

When this input impedance diagram is inverted, the well known circle diagram for the induction motor is obtained as shown in Figure 2.3. This diagram is a good approximation for the normal case of sinusoidal current waveforms and assumes that the air-gap flux is constant at all speeds and

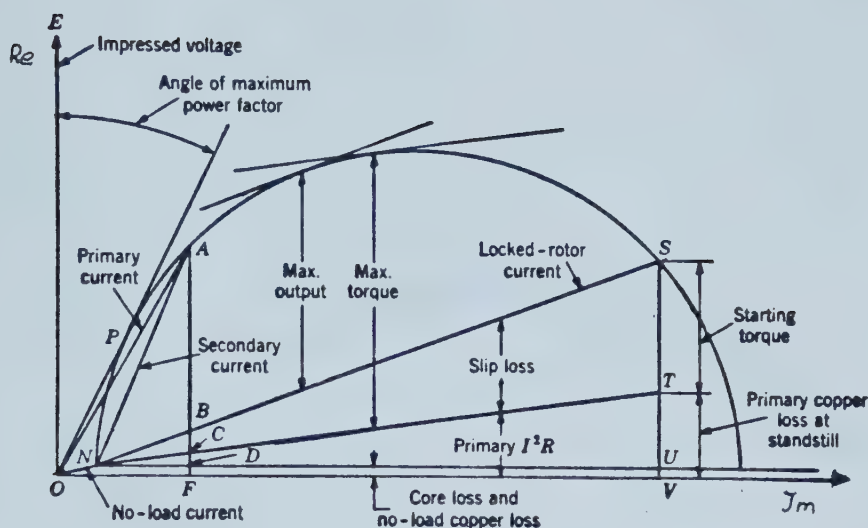


Figure 2.3 Current as a function of slip [5].

the frequency is low enough that no skin effect occurs. In the inverter fed case, however, currents are nonsinusoidal, therefore harmonic losses and torques have to be taken into account and the circle diagram changes with frequency. In the varying frequency case, the reactances decrease linearly with decreasing frequency, but the winding resistances remain constant. The current locus for the variable frequency case can be obtained by impedance inversion again as in the constant frequency case.

Figure 2.4 shows the current loci for the varying frequency case with constant Volts/Hz operation. The impedance circle (its size is determined by X_{22}/X_m^2) reduces in size and moves upwards as frequency decreases. The torque-speed characteristic, as shown in Figure 2.5, can be obtained from the circle diagram of Figure 2.2, by taking

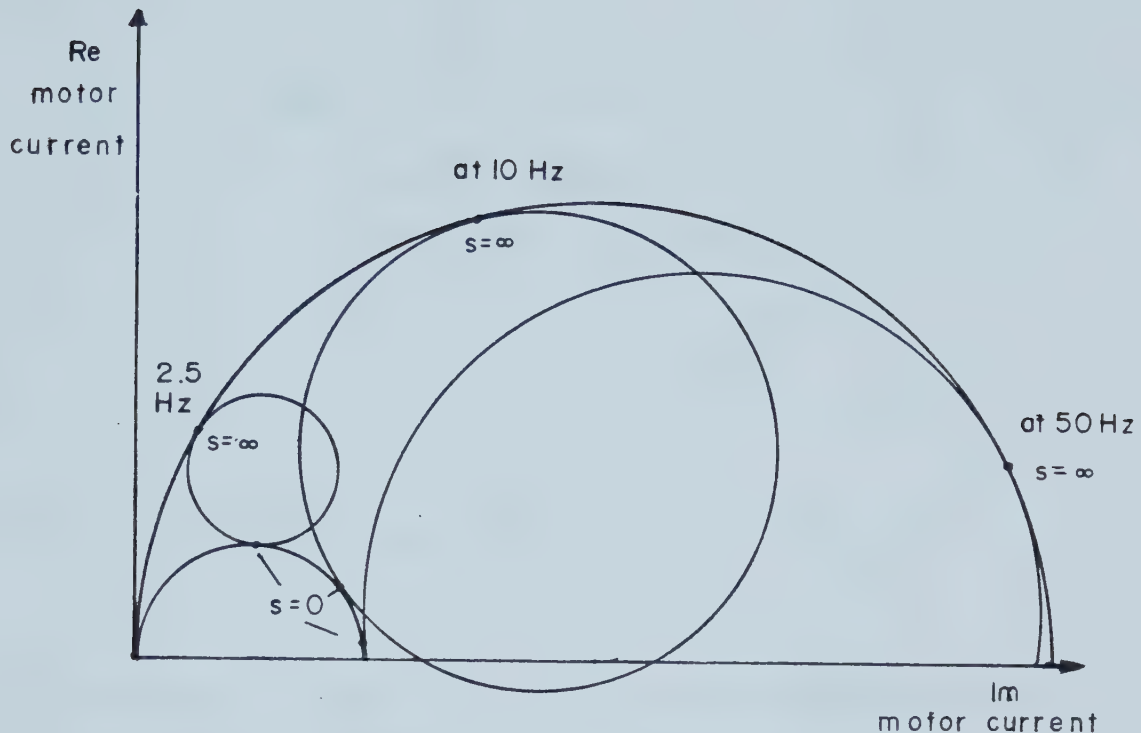


Figure 2.4 Circle diagram for varying frequency [2].

the length of line AC for different values of slip. Figure 2.5 shows that at rated load the machine is rather "stiff". The variation of speed is small for varying load torques. When the loading torque is increased to the peak value, the machine becomes unstable, since the produced torque cannot balance the load-torque. Furthermore this shows that the rated torque is about only $1/3$ of the peak value in order to maintain high efficiency and stability. The efficiency is slightly less than $1-s$. It is apparent that a small slip value at rated load is needed. Rated slip is about 5 % and pullout slip (the slip for pullout torque)

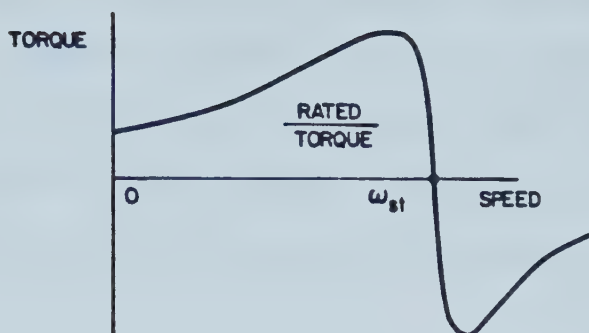


Figure 2.5 Torque-speed characteristic [3].

is about 20 %. Hence the normal operating range for an induction motor yields an efficiency of 85-95 % at a maximum power factor of 0.85-0.9, lagging. The controlled speed drive, however, allows operation closer to the maximum torque without loss of stability if the rotor resistance is reduced, as a design feature, in order to keep rotor losses small.

2.2 Inverter Types

Variable speed ac-drives require a three-phase, variable frequency voltage or current source. To achieve this, several methods can be employed. The most practical ones are described briefly and their advantages and disadvantages stated.

2.2.1 The Cycloconverter

This type of variable frequency, variable voltage converter is shown in Figure 2.6. For simplicity, only the single phase version has been drawn, but the three-phase cycloconverter is comprised of three identical circuits similar to the one illustrated in Figure 2.6.

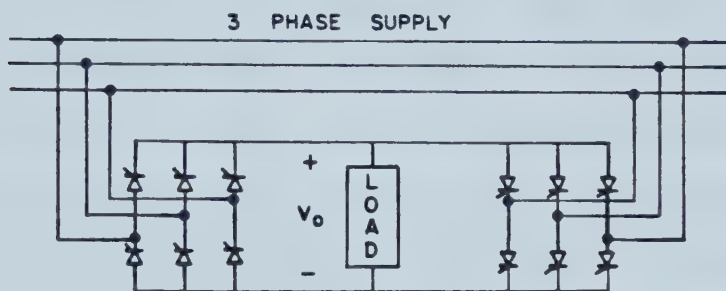


Figure 2.6 Cycloconverter [3].

The single phase cycloconverter comprises two 3-phase bridges connected together back-to-back. The output voltage of the converter consists of segments of the input sine wave which, in a highly inductive load, generates a sinusoidal current with low ripple component.

Advantages of the cycloconverter include:

- natural commutation,
- generation of sinusoidal currents,
- intrinsically high power output, because of the large number of thyristors involved, and

- operation down to zero frequency.

Disadvantages include:

- poor power factor,
- high demand on control and gating circuitry and
- reasonable operation only up to half of the supply frequency

The cycloconverter is a high-power, low-frequency ac-source and is being applied in rolling mills, steel factories and other cases where low speed but high power is required. Because of the high number of thyristors involved, this type is not suitable for microcomputer controlled ac drives, because the micro would be so busy generating pulses, that there would be no time left for current control or other calculations. Furthermore, the limitation on output frequency is not acceptable for many applications.

2.2.2 The Current Source Inverter

The current source inverter used for variable speed drives is shown in Figure 2.7. It requires the least expense for electronic components, although it makes use of large capacitors and has other disadvantages. The current source inverter requires a large reactor as a dc-choke. A controlled rectifier feeds the dc into the link, while the inverter switches this current sequentially to the phase windings. The output current waveshape is approximately rectangular. The capacitors connected to the thyristors in the inverter part, serve as commutating devices and have to

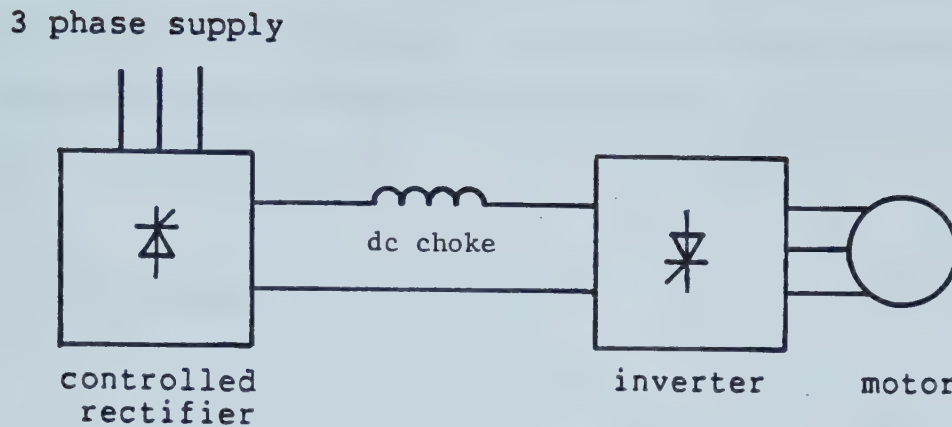


Figure 2.7 Current source inverter

carry the full load current at commutation instants.

Advantages of this system include :

- simplicity of operation,
- regenerative braking without additional circuitry and
- rectifier type thyristors (slow recovery) can be used, because of large commutation margin.

Disadvantages include :

- poor power factor at low speed and light load,
- rather high losses in the semiconductors. (6 junctions in series in each conducting cycle. Rectifier-grade thyristors have slow recovery and higher losses),
- the dc-choke must be large,
- very low speed not possible because of torque pulsations,
- switching losses increase with output frequency, and
- output voltages show large spikes at commutation instants

which produce harmonic torques that interact with the fundamental of the current. Because of these voltage transients, semiconductors with higher voltage ratings have to be chosen.

2.2.3 The Voltage Source Inverter

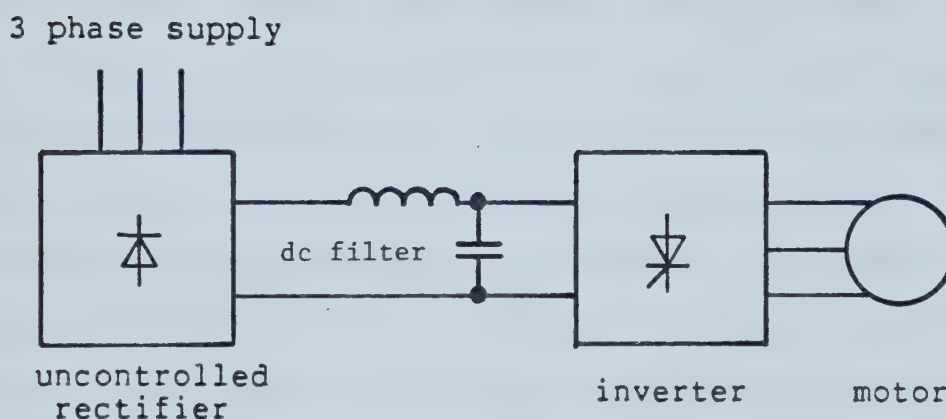


Figure 2.8 Voltage source inverter

The voltage source inverter operates with a fixed dc-bus voltage which is provided by an uncontrolled rectifier bridge through a dc-filter. The uncontrolled bridge eliminates displacement reactive power so that a good power factor is achieved (0.955 theoretically).

Advantages of the voltage source inverter are:

- high frequency operation,
- excellent power factor and
- thyristor and motor insulation requirements not increased

over normal rating.

Disadvantages include :

- extra commutation thyristors needed and
- very low speed operation not possible because of pulsating torques.

2.2.4 Pulse Width Modulator

The pulse width modulator can be used for both types, current source inverter and voltage source inverter, though with constant voltage operation it shows the best results. The operating principle is to approximate a sinusoidal output current waveform by very fast switching of rectangular blocks. The harmonic content is substantial, but occurs at the high switching frequency so that there is little or no interaction with the fundamental current component. This means that only small pulsating torques will be present. One possible output voltage of a pulse width modulator and the corresponding current are shown in Figure 2.9.

This method has the following advantages:

- sinusoidal current(no pulsating torques) and
- operation down to very low speed

Disadvantages of the system include :

- more complex control strategy and
- at higher output frequencies, the commutation losses become

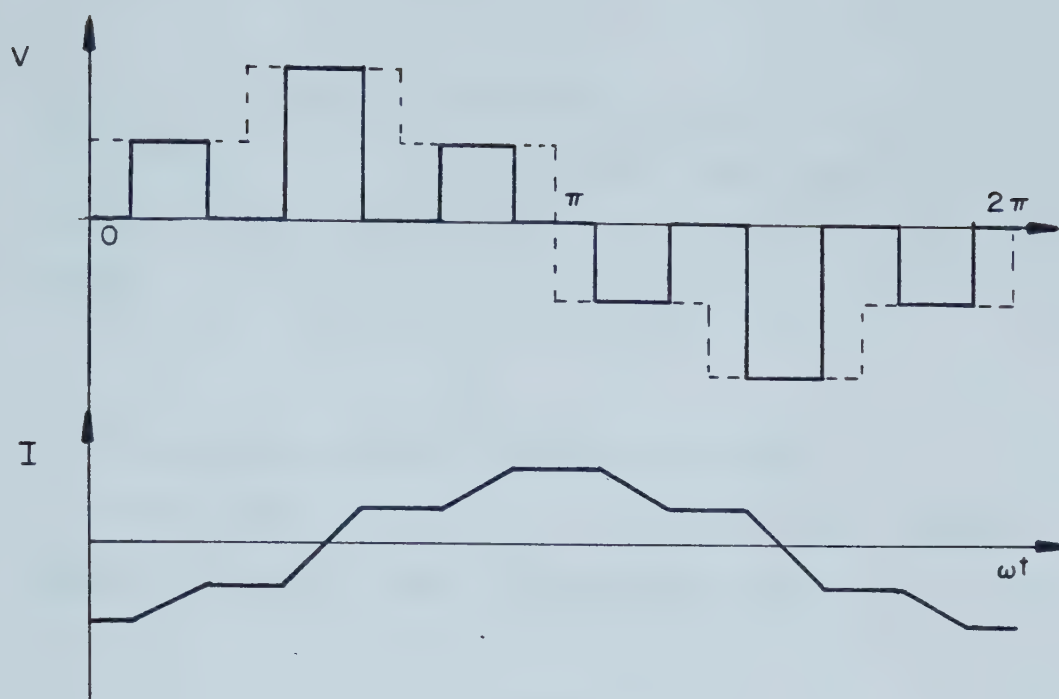


Figure 2.9 Output signal of the pulse width modulator

very high, thus reducing the efficiency.

It has been shown in the previous lines, that the voltage source inverter has good high frequency performance and the pulse width modulator has better low frequency performance. Since the difference of both inverter types is only determined by the application of different firing pulses whereas requirements on power circuits and commutation circuitry are the same, the perfect solution for a variable speed drive is the combination of both types. With a microcomputer that generates the gating pulses for the inverter, the transition from pulse width to six step operation is very easily achieved. The number of pulses per cycle is limited to about twelve, for frequencies in the range from 1 to 30 Hz because of the signal processing speed of the computer. This statement is true, if all the timing

calculations and the generation of pulses is done by the microcomputer itself, but becomes less important if the pulses are generated by external hardware, like programmable timers.

2.3 Control Methods for Induction Motors

The output torque of an induction motor depends on the supply voltage as shown in equation 2.2

$$T = K_T V^2 \quad (2.2)$$

The torque-speed characteristic for different supply voltages is shown in Figure 2.10.

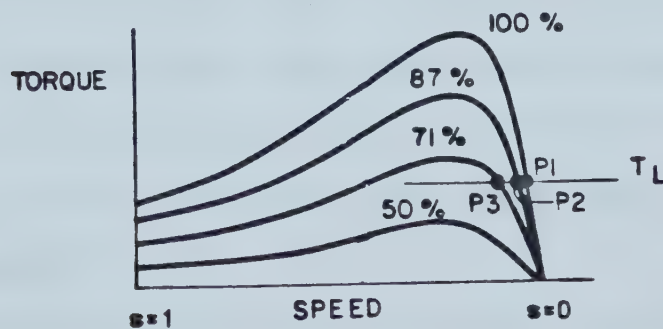


Figure 2.10 Torque speed characteristic for voltage variation

As noted in Figure 2.10, the simplest method of

speed-control would be to vary the stator voltage. This method gives poor speed-control and has slow response to load changes. Furthermore, the slip rises to large values, resulting in increased motor heating since

$$P_{\text{loss}} = P_{\text{el}} \cdot s \quad (2.3)$$

The fact that the motor torque also decreases with decreased stator voltage means that a wide range speed variation is impossible using this method.

Better methods of speed control are available and they all employ inverters of one of the types described previously.

One method is to maintain a constant stator current into the induction motor even in transient states, for instance during a sudden load increase. This strategy requires that the converter respond over the entire speed-torque range otherwise there would be stability problems. But this constraint limits the speed range over which the strategy can be applied.

A second method of speed control is to calculate the slip frequency using measured values of terminal voltage and current. In order to calculate the slip by this method one has to know the induction motor parameters accurately. Since the resistances vary widely with temperature, an accurate model is difficult to achieve.

A. B. Plunkett in [9] describes a method of controlling the phase angle between motor current and motor flux. When determining the position of the flux vector, one either has to compute it from measured current and terminal voltage, with the disadvantages described above, or to sense it directly by introducing a Hall-probe into the air gap of the motor. This method may be suitable for laboratory use, but because of its lack of robustness, the Hall-probe would be destroyed in a few weeks in a field application.

A very simple method is to sense the shaft speed and control the terminal voltage so that the ratio of voltage/frequency (V/f) is kept constant. The ratio V/f is a good approximation for the airgap flux at medium speed ranges so the steady state performance is very good at these frequencies. As described earlier, the influence of the resistance at low frequency is increased over the influence of inductances so V/f is not representing the motor flux anymore and at high speed one runs into voltage limitation problems.

The phenomenon of reduced flux at low speed will be explained further in the following section. Recall that the equivalent circuit for the induction motor was shown in Figure 2.1. The following phasor equations may be derived from this diagram

$$V_1 = (R_1 + jX_1) I_1 + \left(\frac{R_2}{s} + jX_2 \right) I_2 \quad (2.4)$$

$$jX_m(I_1 - I_2) = \left(\frac{R_2}{s} + jX_2\right)I_2 \quad (2.5)$$

where the slip is defined as

$$s = \frac{f_2}{f_1} \quad (2.6)$$

The index 2 always refers to the rotor side and the index 1 to the stator side when dealing with induction motors.

The motor torque can be derived as,

$$T = \frac{p_m}{2\pi f_1} (I_2)^2 \frac{R_2}{s} \quad (2.7)$$

and inserting results from eq (2.5) and (2.6) yields,

$$T = \frac{p_m}{2\pi} \left[\frac{V_1}{f_1} \right]^2 \frac{f_2 X_m^2 / R_2}{\left[R_1 + \frac{f_2}{f_1 R_2} (X_m^2 - X_{11} X_{22}) \right]^2 + \left[X_{11} + \frac{f_2 R_1 X_{22}}{f_1 R_2} \right]^2} \quad (2.8)$$

The torque changes with the square of the flux and the flux is proportional to the ratio V_1/f_1 . More precisely it is E_1/f_1 , but usually the voltage drop across R_1 and X_1 is negligible. This is not true for low frequencies. The voltage drop across R_1 is the same for all frequencies at rated load, but at low frequencies this is a considerably higher fraction of the supply voltage than at rated frequency.

The remedy for increasing motor performance is a voltage boost, which may then give constant air-gap flux at all frequencies. A truly constant flux, however, can only be

obtained with a feedback system that measures the actual flux by means of some sensor (coil or Hall-probe). The sensor makes the system more expensive and liable to failures, hence an open loop system will be preferred, although this may lead to a higher core saturation at light loads. The core losses are not seriously increased by such a system, because of the low stator frequencies.

The control strategy that shows the best results is the constant slip method. The torque-speed characteristics for slip control are shown in Figure 2.11. From this picture it is seen that the motor develops a constant torque up to rated frequency. As frequency increases further, the torque developed varies inversely with frequency but the power delivered to the load remains constant. This resembles the torque-speed characteristic of a d-c shunt motor. The constant slip control combines high efficiency of the motor operation with excellent speed variation.

The advantages are of this system are :

- motor operation at rated torque over the whole speed range,
- high power-factor over wide speed range, and
- by simply changing the control parameters this configuration can be adjusted to a variety of applications

For traction applications a large starting torque is needed with less torque required at higher speeds. This is exactly the characteristic of the curves in Figure 2.10. Constant torque, constant horsepower, or constant speed

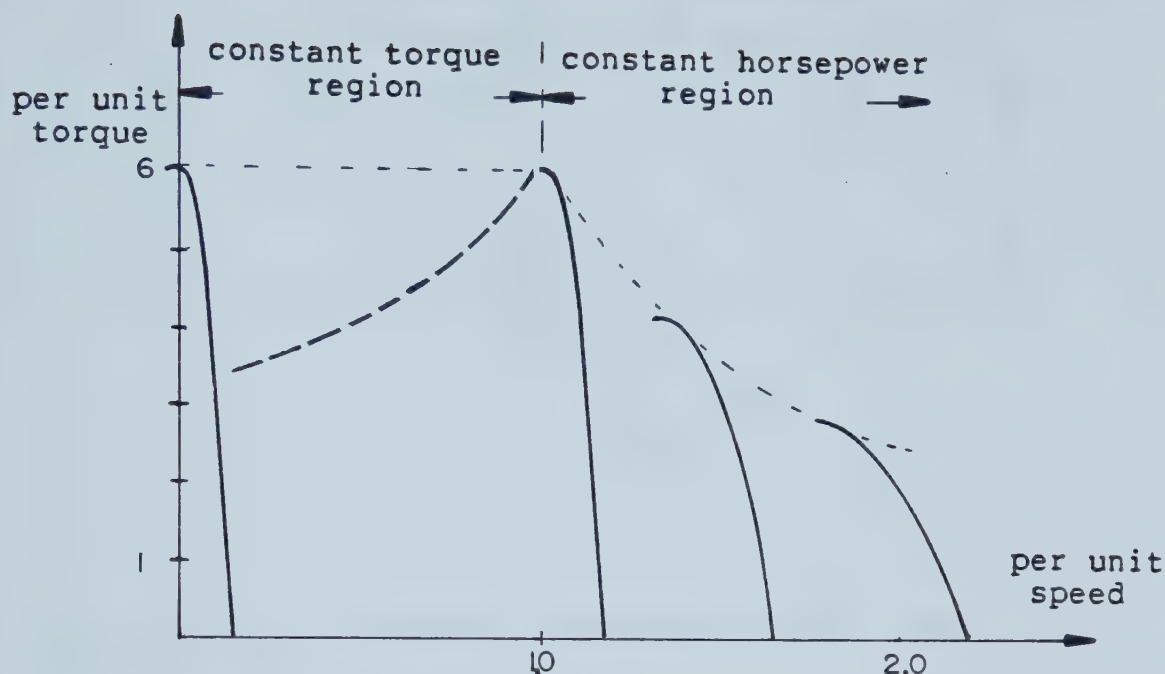


Figure 2.11 Torque speed characteristics with slip control

characteristics can also be obtained with the controlled slip drive.

In the microcomputer implementation of slip control, shown in Figure 2.12 a speed sensor (for example an optical tachometer or magnetic pickup) measures the actual rotor speed and the microcomputer compares this to the desired speed, which is obtained by subtracting the given slip value from the stator frequency. The stator frequency then is adjusted by changing the gating frequency of the three-phase inverter until the desired rotor frequency is reached. The speed control algorithm must have a slip-limit of less than the breakdown slip in order to not exceed the breakdown torque and to stay in the stable region of the torque-slip curve. This is relevant for both, motor and generator

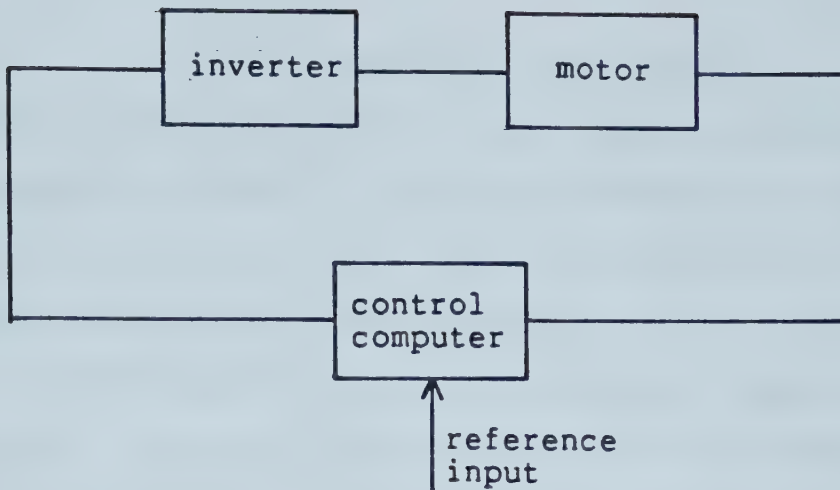


Figure 2.12 Microcomputer implementation of the controlled slip drive

operation. A current limiting function also has to be incorporated since the control strategy does not inherently limit the motor current.

3. Description of the Inverter

Two major advantages of the dc-side commutated inverter, illustrated in Figure 3.1, were instrumental in selecting it as a means of speed control. Firstly, the dc-side commutated inverter allows rather high switching frequencies. The upper limit is determined by the reverse-recovery time of the thyristors. The second advantage is that the commutation circuit consisting of thyristors ThA to ThD, L₁, L₂, and C₁, limits the rate of rise of reapplied forward voltage (dv/dt) to moderate values. This makes the use of snubber circuits unnecessary and therefore inverter losses are reduced.

3.1 The DC-Side Commutated Inverter Operation

The dc-side commutated inverter operates with a fixed dc-bus voltage, the output voltage magnitude being modified by varying the time-ratio of the conduction interval.

Thyristors Th₁ to Th₆ make up the three phase inverter that alternately connects the motor windings to the positive or negative dc-bus terminal. In order to produce an output-phase rotation ABC, three thyristors always have to be turned on in a conduction interval. Two of these in either the upper or lower half of the bridge, the third in the opposite half. From the line diagram of Figure 3.2, it can be seen that the conduction interval of any thyristor is 180° as opposed to 120° in a dc-drive thyristor bridge. The freewheeling diodes D₁ to D₆ of Figure 3.1 allow any

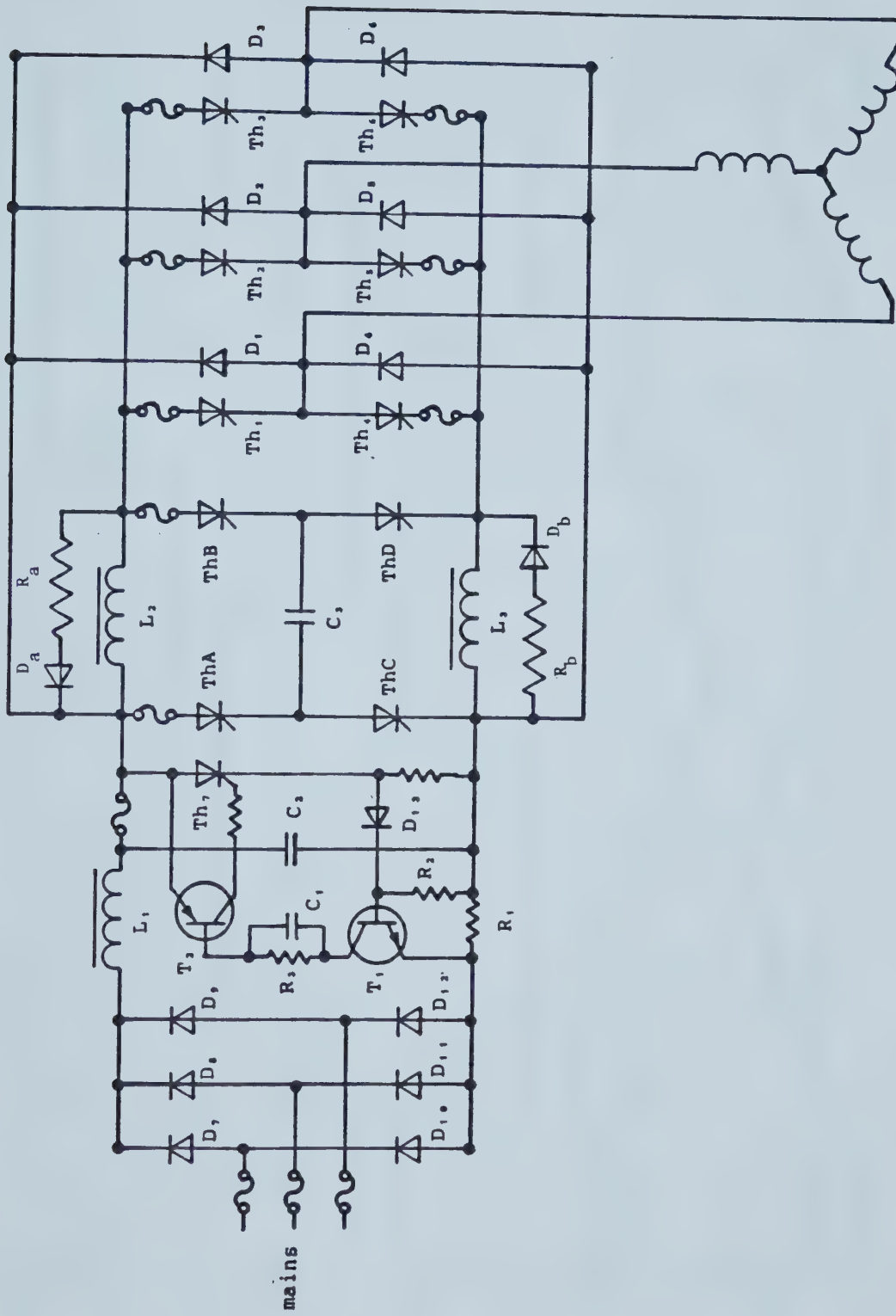


Figure 3.1 Schematic of the dc-side commutated inverter

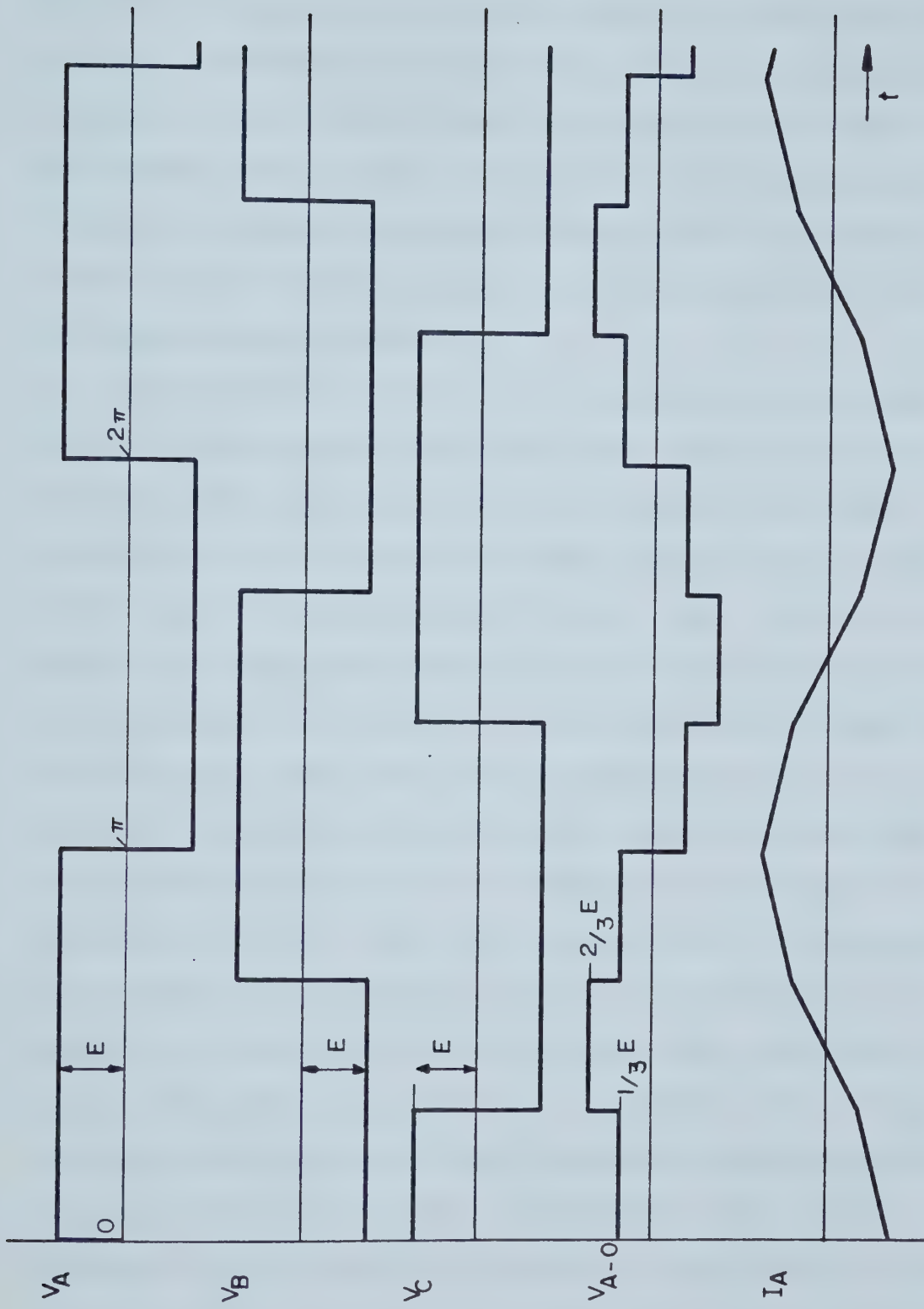


Figure 3.2 Phase voltages and current waveforms for the six step inverter

reactive component of the motor current to circulate when the main thyristors are turned off. Whenever a commutation sequence is performed, either the upper or the lower half of the bridge is turned off. Since turning off one half of the bridge disconnects the motor from the supply potential, a change of the firing sequence can now be initiated by firing the appropriate thyristors.

The inverter configuration, selected for the study, allows motor operation only and very little regenerative braking. Since the rectifier is an uncontrolled diode bridge, it is not capable of feeding power back to the three phase supply like a thyristor bridge would. Therefore, any amount of energy generated by braking the motor only serves to charge the filter capacitor and the excess voltage has to be limited by some means. This is done by connecting a resistor in parallel to the filter capacitor. Hence the efficiency of the inverter is reduced. If regenerative braking is desired, some overvoltage control circuit should be employed. A suitable way of doing this would be to connect a resistor and a high-voltage power transistor in series and connect this circuit in parallel to C_f . The transistor would be switched on whenever the voltage exceeds an upper limit and switched off slightly above rated voltage. The transistor and resistor would have to be able to carry more than twice the rated motor current for short periods to make this limiting circuit effective.

3.2 The Commutation Sequence

The commutation sequence has to be explained more thoroughly since an understanding of this concept is essential for safe operation, trouble shooting and for understanding the design requirements of Chapter 3.5. Assuming the commutation capacitor C_c of Figure 3.1 is charged to the supply voltage E with the left-hand plate positively charged from the previous cycle, the commutation is initiated by gating thyristors ThB and ThC simultaneously. This connects the capacitor C_c to the positive inverter bus. The voltage across the upper conducting thyristors is reversed to $-E$ and the upper commutation inductor L_c is forced to support a voltage of $2E$. The negative voltage across the conducting thyristors forces them to go off. The load current flowing in L_c is diverted to C_c and charges it up to the positive supply voltage. At the instant C_c is charged to $+E$, the free-wheeling diode DA starts conducting and takes over any current flowing previously through C_c . This causes ThB and ThC to go off. The path for reactive components of the load current is provided by the lower diodes and that thyristor in the lower half-bridge which had been conducting prior to commutation. This completes the positive commutation sequence.

The procedure for commutation of the lower half is the same, except for ThA and ThD now being gated on. The commutation circuit and its simplified equivalent are shown

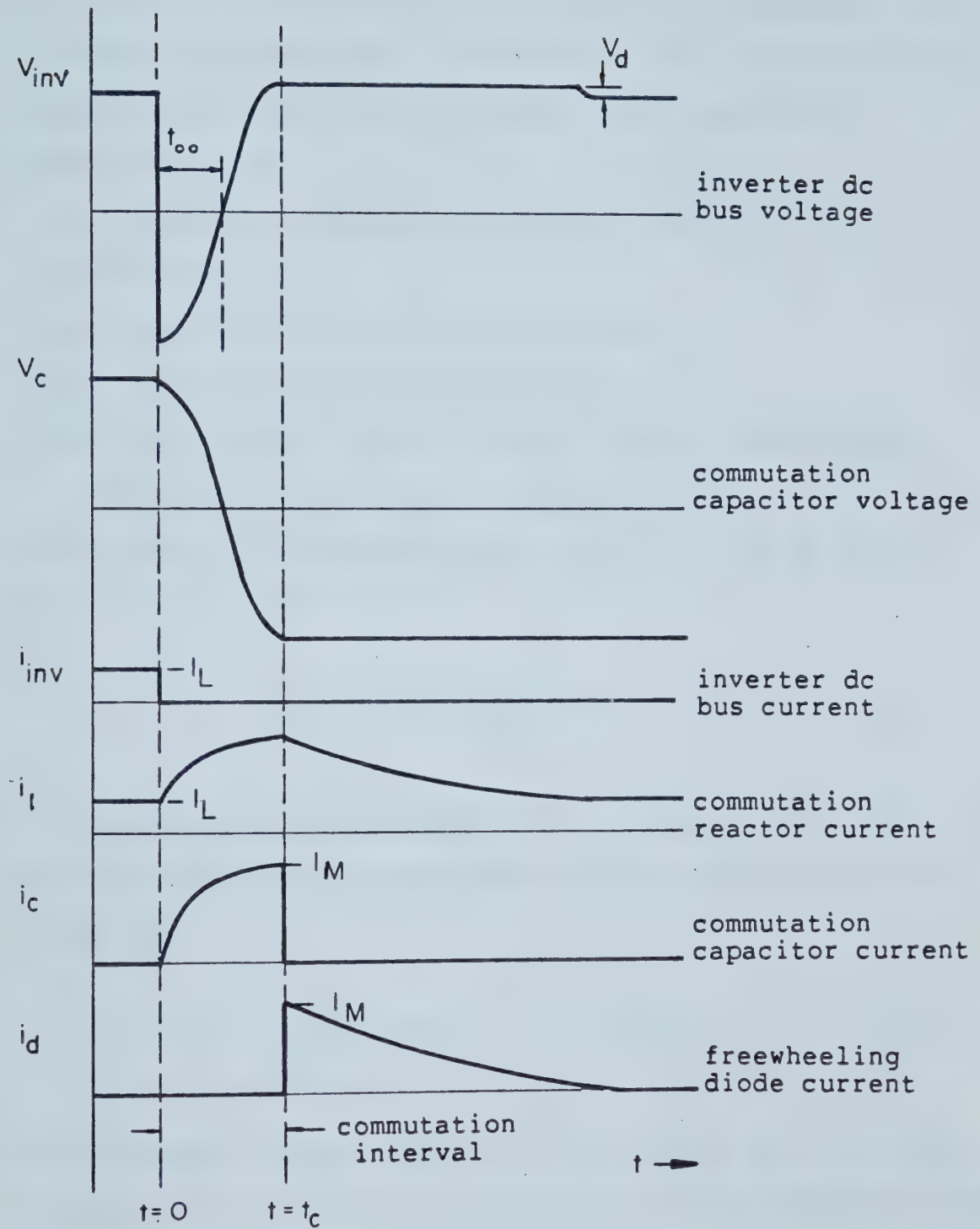


Figure 3.3 The voltages and currents at commutation

in Figure 3.4. The circuit is a simple LC circuit and under a number of assumptions, its behavior can be described by a second order differential equation. The simplifying assumptions are:

- the inductance is concentrated in L , (or L , respectively),
- the capacitor voltage has no overshoot,
- the thyristors are perfect switches, and
- the load current remains constant during commutation.

The differential equation describing the circuit during this portion of the commutation cycle is given by Equation 3.1

$$L \frac{di_2}{dt} - e_c(0) + \frac{1}{C} \int i_2 dt - E = 0 \quad (3.1)$$

with the initial conditions $i_2(0+) = I_L$ and $e_c(0+) = -E$.

Solving this equation with the initial conditions inserted leads to,

$$i_2(t) = -\frac{2E}{\omega L} \sin \omega t + I_L \cos \omega t \quad (3.2)$$

This solution is valid until C , has charged up to $+E$. The commutation time-interval t_c is found by differentiating Eqn (3.1) and letting $di/dt = 0$.

$$\frac{di_2}{dt} = \frac{2E}{L} \cos \omega_0 t_c - I_L \sin \omega_0 t_c = 0 \quad (3.3)$$

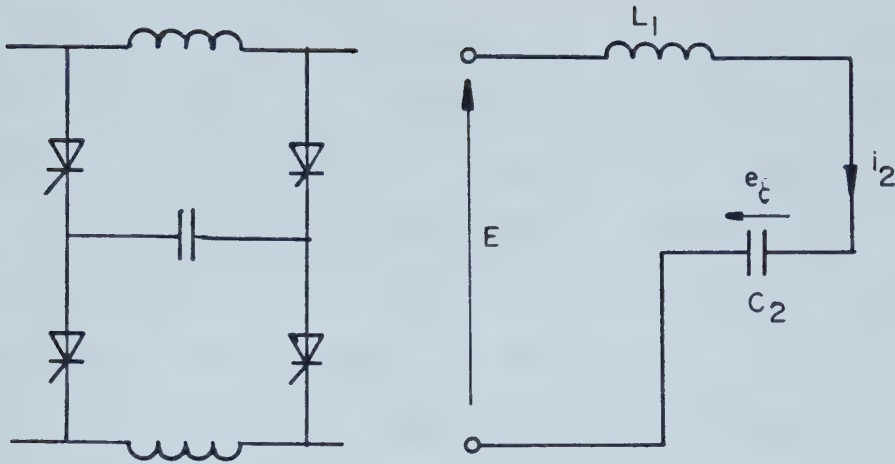


Figure 3.4 Equivalent circuit at commutation instant

The current in the reactor L_1 has reached its maximum at the end of commutation and its magnitude can be calculated by substituting t in Equation 3.2.

$$I_M = \sqrt{\frac{4 E}{\omega^2 L^2}} + I_L^2 \quad (3.4)$$

In order to minimize the commutation losses of the inverter, the trapped energy in the reactor has to be minimized. An iteration method to obtain minimum trapped energy is described in reference [14]. Using the results of reference [14], a general expression for computation of the values for L_2 , C_2 and C_3 , can be obtained. They are given by

$$L = \frac{1.82 E t_{00}}{I_L} \quad (3.5)$$

$$C = \frac{1.47 I_L t_{00}}{E} \quad (3.6)$$

where t_{00} is the thyristor turnoff time.

As mentioned above, the commutation circuit determines the dv/dt for the main thyristors, thus eliminating the need

for snubber circuits to protect the thyristors. However, this does not apply for commutation thyristors ThA to ThD. Consider the case where the capacitor has been charged with the left-hand plate positive. Before commutation, all 4 thyristors are off and there is no potential difference between anode and cathode for ThA and ThD. Now ThB and ThC are gated on and the cathode of ThA drops suddenly (in about $2\mu\text{s}$) to the negative bus potential. If the supply voltage is high enough, the dv/dt will exceed the rated value for the chosen thyristor and will cause a short-circuit across the dc-bus, since ThA turns on when ThD is already conducting.

The situation for ThD is different. The right-hand plate of C_s was charged negatively, but after ThB is gated on, there is still no voltage difference from anode to cathode, so this thyristor will not turn on. When the second pair of thyristors is gated on, thyristor ThC is now subjected to the high dv/dt . So the left hand branch, which has the least protection by other circuit elements, is always subjected to the high dv/dt .

A second threat to the life of the semiconductors is caused by their own turn-on delay time. When ThB and ThC are gated on, ThB will start conducting after about $2\mu\text{s}$ delay. If ThC did not begin to conduct, there is an initial voltage step across it with the magnitude of E that has a very fast rate of change. The blocking voltage of ThC is therefore $2E$. The way to overcome these problems is to insert passive elements in the commutation circuitry. Snubber circuits

across each thyristor are used to reduce the voltage spikes and assure equal voltage distribution in every branch. A small inductance in series with the commutating capacitor gives satisfactory results. The additional inductance is small compared to the size of the commutation inductor L , or L_1 , so the commutation time is not affected, but the dv/dt across the nonconducting thyristors is decreased. The equivalent circuit for one thyristor can be seen from Figure 3.5.

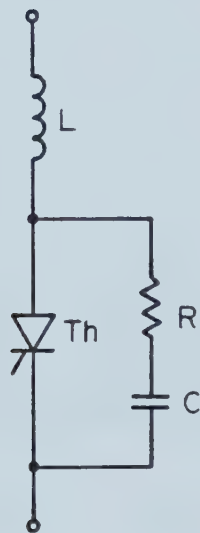


Figure 3.5 Thyristor with snubber circuit

The voltage spike produced by the mechanism described lasts for about $1\mu s$. To filter this spike, the time constant RC is made about 10 times this value. To determine R , one has to take into account that the energy stored in C will be dissipated in R and Th upon gating the thyristor. The capacitor discharge current will add to the load current I_L and therefore affect the current rating of the semiconductor

device. The discharge current has its maximum value immediately after the turn on of the thyristor. For reliability reasons, the resistor should be chosen to limit the current to $1/4$ of the current rating.

3.3 The Firing Circuit

The main thyristors are fired by optically coupled auxilliary thyristors. The firing circuits comprise only a few components as shown in the circuit diagram of Figure 3.6.

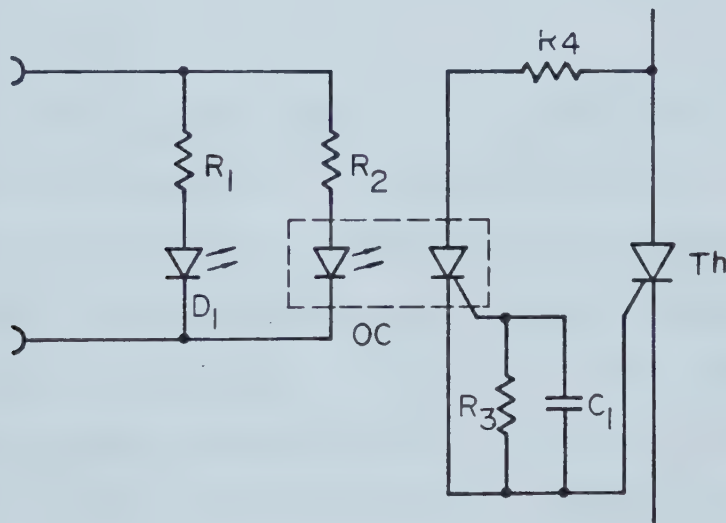


Figure 3.6 Optically coupled firing circuit

In Figure 3.6, D_1 is an indicator LED on the panel of the inverter. The current for this LED is limited to 15 mA by R_1 . R_2 is designed to limit the LED-current of the optocoupler to 40 mA in order to turn on the detecting thyristor sufficiently fast. The C_1, R_3 parallel combination

is for improving the dv/dt characteristics of the optocoupler. R_4 limits the gate current for the main thyristor to 160 mA. This high value is necessary to turn the thyristor on fast enough. There is, however, not much gate power to be dissipated, since the duration of these pulses is only 1 to 5 μs depending on the turn-on delay of the main thyristor. The duration of the pulse that drives the LED was set to 100 μs in order to get all thyristors fired. This duration can be varied by changing the software of the control computer, but must not be made less than 60 μs .

3.4 The Protection circuit

During the testing phase in the lab, and later on in field service, overcurrent conditions may occur. This may result in a commutation failure and subsequently in a short-circuit across the dc-bus. To avoid damage to the semiconductors, they have to be disconnected from the filter capacitor by some hardware device. If this is not done sufficiently fast, the large amount of energy stored in the filter capacitor will be dissipated in the semiconductors that caused the short and will inevitably destroy the semiconductors. Software detection of a commutation failure was considered impractical because a/d-converters would be too slow and blocking of firing pulses would have no effect on thyristors that are already in the conducting stage. As a result of the above considerations a protection circuit, as

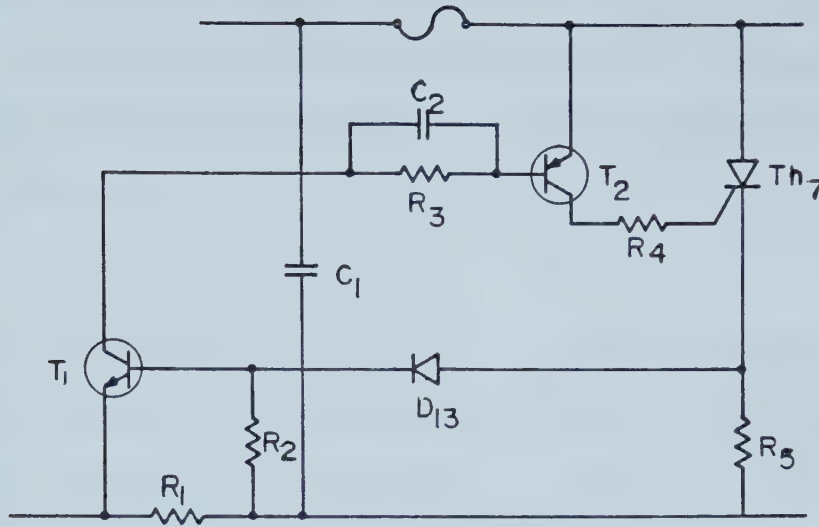


Figure 3.7 Overcurrent protection circuit

depicted in Figure 3.7, was designed. This circuit operates in the following way. Resistor R_1 acts as a current sensor which leaves transistor T_1 in the off-state as long as the dc-current is less than 12 A (the current rating of the chosen thyristors). The position of R_1 was chosen in such a way that the spikes caused by the commutation current do not affect the sensing transistor. This may have a delaying influence on the protection circuit, but the trigger level was set sufficiently low to ensure safe operation in the normal state. As soon as the voltage drop across R_1 exceeds 0.6V, T_1 will start conducting. This ties the base potential of T_2 more negative and this transistor starts conducting also. Thus Th_7 is gated on and eventually destroys the fuse, thus disconnecting the rectifier and filter network from the inverter side of the dc-bus. The combination R_2 and D_{13} have

a positive feedback effect on T_1 . The voltage drop across R_1 (which serves as a current limiter for T_{h1}) drives T_1 (and T_2) completely into saturation to ensure sufficient gate current for T_{h1} .

3.5 Design Considerations for the Inverter Components

The inverter is to be operated from a three-phase ac-supply with the phase voltage $V = 120V$. The output voltage of the three-phase rectifier can be calculated from,

$$V_{out} = \frac{p}{2\pi} \int_0^{2\pi} V_{ph} \sin \omega t \, d\omega t = \frac{6}{\pi} \sqrt{2} V_{ph} \frac{1}{2} = 1.35 V_{ph} \quad (3.7)$$

with $p = 6$ for the bridge rectifier. The motor is rated at,

$$208 \text{ V} / 2.8 \text{ A} \quad (3.8)$$

so the peak current is,

$$2.8 \sqrt{2} \text{ A} = 3.9 \text{ A} \quad (3.9)$$

An increase in current can be expected when the motor is not operated from a sinusoidal supply. The harmonic currents were expected to add about 20% of rated current to the motor current. For overload conditions a current of 6 times rated current has to be assumed, so the thyristors have to be able to carry 12A rms current. Since every thyristor conducts for 180° the average current is $24/2=12A$.

The blocking voltage was calculated to be,

$$V_b = E * 2 * 1.1 * 1.1 = 392 \text{ V} \quad (3.10)$$

with the safety margins determined as follows:

- factor of 2 , since the thyristors may have to withstand twice the dc-bus voltage,
- factor 1.1 for 10% overshoot,
- factor 1.1 as a general safety margin

A blocking voltage of 400V was considered suitable but 600V thyristors were readily available and hence they were used. Because of the higher switching frequencies in the PWM-inverter, fast thyristors had to be employed. Pertinent data of the thyristor is :

$$600 \text{ V} / 12 \text{ A} / 15 \text{ to } 35 \text{ } \mu\text{s}$$

The rectifier diodes each have to carry the load current for 1/3 of the period of the mains voltage , so the average diode current for an overload current of 24 A is 8 A. The commutation elements were designed using Equation (3.5) and (3.6)

$$L = 6.18 \text{ mH}$$

$$C = 1.47 \text{ } \mu\text{F}$$

An initial design was done using a much slower thyristor and when faster thyristors were substituted the original inductors were left in the circuit and only the capacitance was adjusted to meet the new design requirements. This meant that the design is suboptimal but the commutation losses were still within reasonable limits.

The final values for the circuit components were $L = 5\text{mH}$, $C = 3.0\mu\text{F}$. As can be seen from the circuit of diagram Figure 3.8, in every semiconductor path there are protective fuses. Only fast-acting fuses were used to protect the main thyristors. A current rating of 8 A provided sufficient protection. For the diode bridge protection slow-acting, 10 A fuses were sufficient.

The inverter, built of the components mentioned in this chapter, was mounted in a 19" rack. A total number of 10 current shunts were included for measurement purposes. The shunts are located in the 6 thyristor paths, 3 motor phases and one is used for measuring the dc-link current. From this latter one, the signal for the isolating current amplifier is taken and, after amplification, is fed into the control computer. The isolation amplifier is designed to give an output signal of 2V per A. The voltage signal for the control computer is taken from the differential amplifier IC₃ of Figure 3.8. This amplifier has a gain of 1/40 in order to produce an output voltage of 10 volts at an input of 400V.

An interface between the microcomputer and the inverter was designed to drive the optocoupled thyristors and the indicator LED's for each channel. This is an emitter-follower type circuit and is designed to deliver 60mA for the main thyristor-couplers and 100mA for the commutation thyristor-couplers. A total of 8 pulse channels is needed from the microcomputer, six for the main thyristors Th, to

TH, and, since the commutation always fires two thyristors at a time, two more channels are needed.

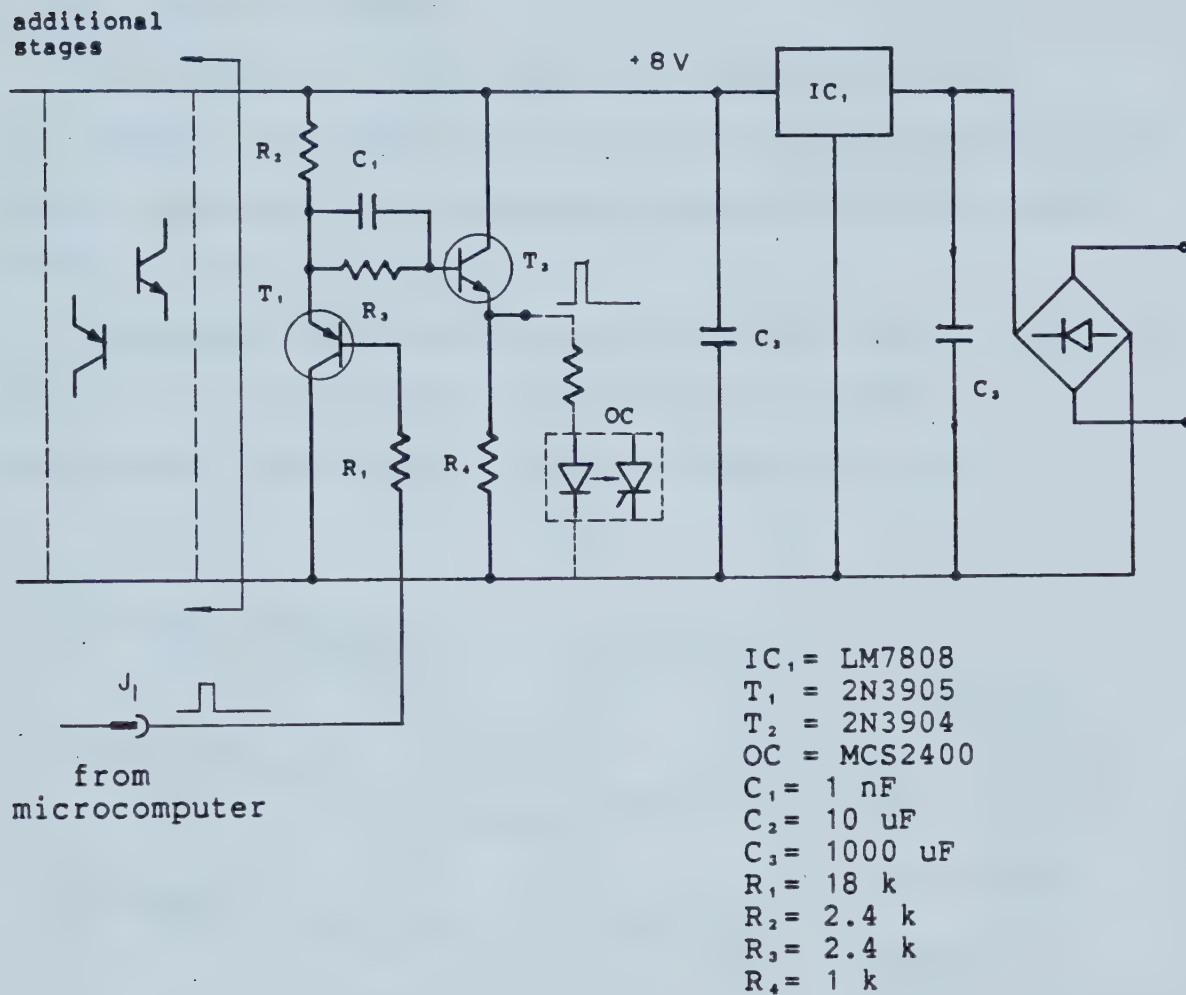


Figure 3.8 The microcomputer to inverter interface circuit

4. Description of the Induction Motor and Related Equipment

4.1 The Induction Motor

The induction motor used in the study was tested thoroughly in the laboratory in order to obtain data for the control algorithm. The parameters obtained from the measurements are given in Table 4.7.

The motor data were measured using the circuit given in Figure 4.1. Various tests were conducted and the experimental results are given in Tables 4.3 to 4.6

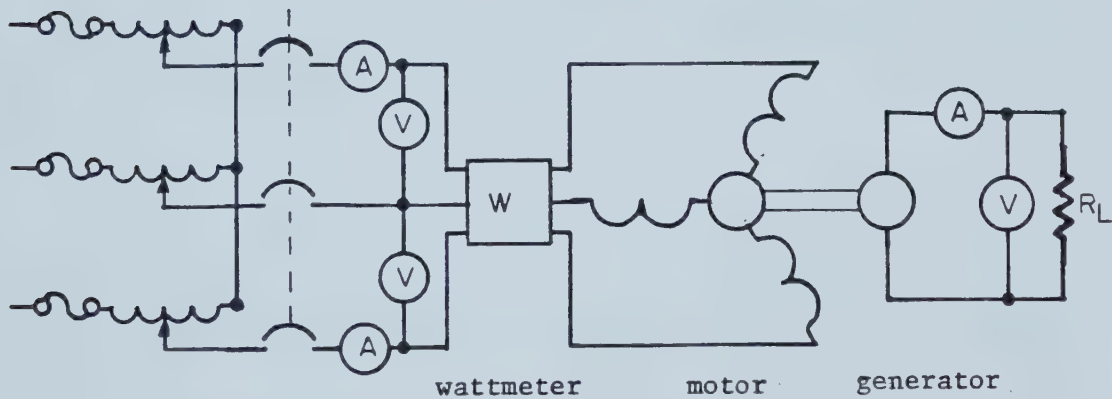


Figure 4.1 Circuit arrangement for measuring induction motor parameters

The current for $s=1$ and rated voltage is computed from

$$I' = \frac{V'}{V} I = 12.22 \text{ A} \quad (4.1)$$

and the total stray inductance from

$$X_1 + a^2 X_2 = \frac{V_{ph}}{I_{ph}} \sin \theta \quad (4.2)$$

Table 4.1 Motor nameplate reading:

Wagner Leland Sangamo Company Ltd. Guelph Can.			
Spec.	912043-01		
	66905		
HP	3/4		
R.P.M	1725		
TYPE	PB 299		
FR	56	CY	60
FORM	BYJ	PH	3
A	2.8/1.4		
V	208-220/440		

Table 4.2 Generator nameplate reading

Leland Electro Guelph Can.			
TYPE	D62		
FR	224	HP	3
FORM	BOVJH	CY	DC
RPM	1725	PHASE	CP
A	12.5	V	230

Table 4.3 Locked rotor test

V	I	W	$\cos \theta$	P
45	2.95	0.08	0.22	50
43	2.9	0.12		

finally the combined stator and rotor resistances from

$$R_1 + a^2 R_2 = \frac{V_{ph}}{I_{ph}} \cos \theta$$

Table 4.4 No load test (s = 0)

V	I	W	cos θ	P
188	1.38	-.15	.177	84
188	1.52	.36		

The results from this test yield the magnitude of the

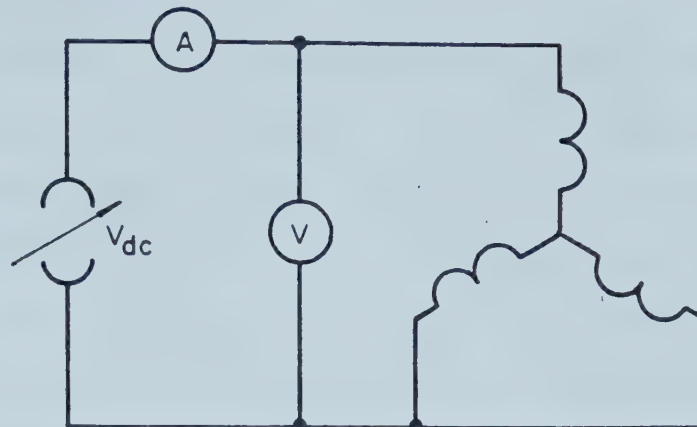
magnetizing inductance $X_m = \frac{1}{\frac{I_{ph}}{V_{ph}} \sin \theta}$ (4.3)

Table 4.5 Load tests

rpm	s	I	W	I	V	cos θ	P	comments
1758	.008	1.75	.32	1.15	140	.762	428	2A/200
		1.7	.75					
1730	.038	2.3	.2	1.8	147	.776	550	5A/200
		2.05	.35					
1720	.044	2.45	.59	2.1	145	.808	612	2A/200
		2.2	.94					
1695	.058	3.15	.33	3.1	147	.868	820	5A/200
		2.7	.49					
1595	.119	4.85	.56	5.4	133	.905	1370	5A/200
		4.45	.81					
1360	.244	6.8	.62	7.6	92	.905	2040	5A/200
		7.05	.4					

Table 4.6 Stator resistance measurement

V	I
15	2.8
10.6	2
7.9	1.5
5.1	1
5.4	1
4.4	.8
3.2	.6
2.2	.4



With these data, all motor parameters were determined with the derived parameters shown in Table 4.7

Table 4.7 Derived motor parameters

R_1	=	3.55	Ohms
$a^2 R_2$	=	2.74	"
X_1	=	3.09	"
$a^2 X_2$	=	3.09	"
X	=	76.73	"
L_1	=	8.2	mH
$a^2 L_2$	=	8.2	"
L	=	0.203	H

4.2 The Control Computer

The microcomputer used was available as part of the Power Lab Equipment. It consists of an LSI11/2 CPU with serial and parallel input-output boards, a real-time clock, a/d and d/a converter boards and two magnetic tape drives, all combined in one enclosure. This will be referred to in the following pages as a SNORKEL-terminal, according to the manufacturers terminology. Programs can either be run in stand-alone mode on SNORKEL or under assistance of a host-computer. The host in this case was an LSI-11/23 which supports higher level languages like FORTRAN 77 and C. The powerful features of the C language, in combination with the debugging facilities of the UNIX operating system, made program development quite easy.

The LSI-11/2 has hardware floating point implementation which is not generally found in microcomputers. This is of importance in certain applications and especially in real-time control of fast processes like motor control. A weakness of this microcomputer, however, is the limited speed with which it can do programmed I/O. This drawback didn't concern this project very much since the gating of thyristors required no handshaking signals.

The normal mode of operation in the program development phase is the UNIX-assisted mode. In this configuration a standard data terminal is connected to SNORKEL which is linked to the host computer by means of a serial line and two high speed parallel lines. The only task for the SNORKEL

terminal to perform in the application software development phase is echoing characters back and forth between the data terminal and the host computer using the serial line. This is called the transparent mode. Once a program is to be executed by SNORKEL, this program is loaded into its memory from the host computer using the parallel lines and, on completion of the loading operation, control is transferred to the SNORKEL. The host then acts as a supervisor and allows the SNORKEL access to its (the hosts) file system and operating system.

In contrast to this is the stand-alone mode where SNORKEL receives commands from the data terminal only. A stand-alone version of a program can be obtained by using a different compiler and loader option than in normal compilation. The stand-alone version can be saved by writing it on magnetic data-tape. This operation makes use of the second tape drive in SNORKEL, in conjunction with the tape handling routines. The tapehandling software package contains routines which make the tape self-booting. Whenever SNORKEL is switched on, it first reads the tape mounted in drive 0 and loads the contents from that tape into memory, this can be either a transparent task or an executable program.

5. Program Description

5.1 Main Routine

After being loaded into the RAM of SNORKEL, either from the host computer or from a magnetic tape, the control program starts by asking the user for input. The main routine is the one that branches to several subroutines according to the information given by the user. The flowchart is shown in Figure 5.1 and a list of commands is given in Table 5.1.

Table 5.1 Commands for the motor control

Catalog	produces a list of command on the terminal
Help	does the same
Display	shows values of voltage,current,stator - frequency, as well as controller gains
New Frequency	asks for input and starts motor at the new speed
Parameters	enables the operator to change control gains on line
Reverse	lets the motor spin backwards
Soft	performs a softstart from zero speed
Torque	Asks for a limit and then increases the speed up to that value hereby keeping the torque in a certain range
Quit	stops execution
Stop	does the same

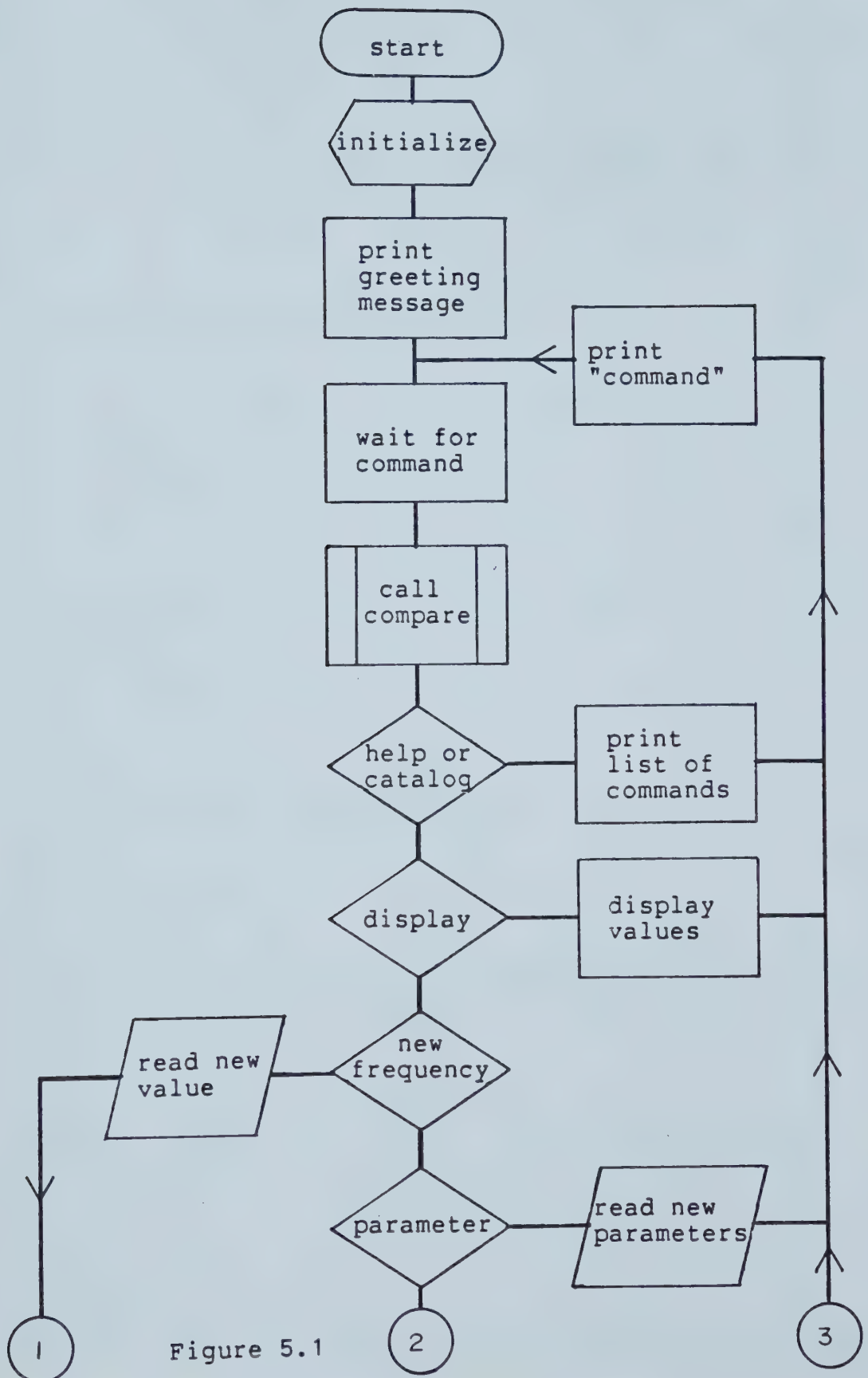


Figure 5.1

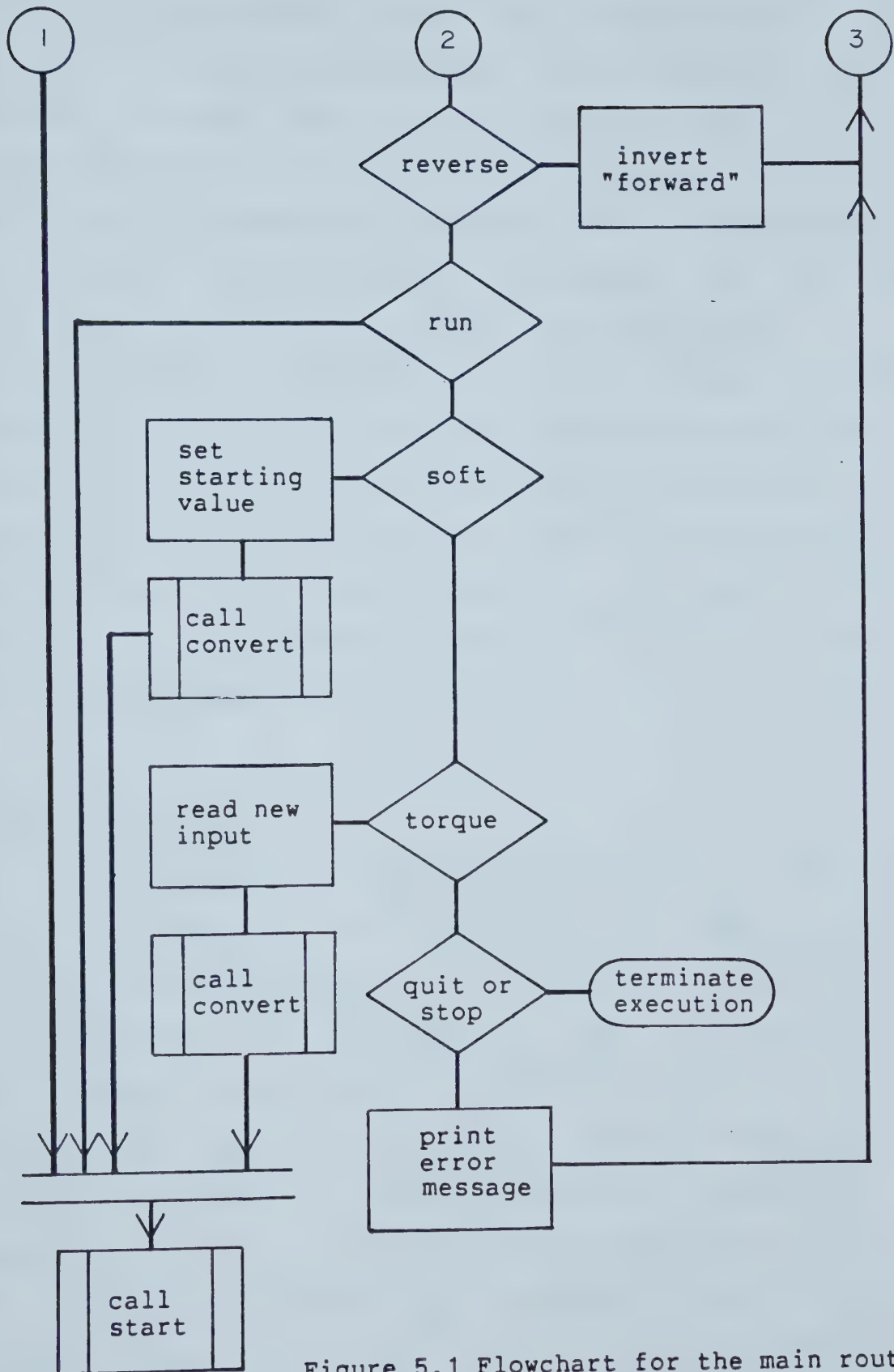


Figure 5.1 Flowchart for the main routine

After initialization of the global variables the routine prints a prompting message on the terminal and then waits for commands. After a keyword has been typed in, a routine "compare" is called that simply compares the keyword with a list of commands and returns a number corresponding to the position of the keyword in the command list. Only the first three characters are used for the comparison to facilitate easy use. The commands in the list were stored in alphabetic order for the search. The number returned by the comparison routine is used for branching to the several functions. The command "reverse" may only be issued after the motor has come to almost standstill. The reason for this being that it is impossible to feed power back to the mains as described in Chapter 4.1

5.2 Routine "start"

If the command "run" is issued, the routine called "start" is entered and the thyristors are switched on in a circular fashion. Referring to the flowchart of Figure 5.1, it is noted that the actual timing is done in an infinite loop in which another loop is embedded. The only exit from that infinite loop is through typing something on the keyboard. The micro will thereafter perform a commutation sequence and return to the main routine to call for input. The second loop is repeated five times, each time writing a different bit pattern to the digital I/O-board and this way

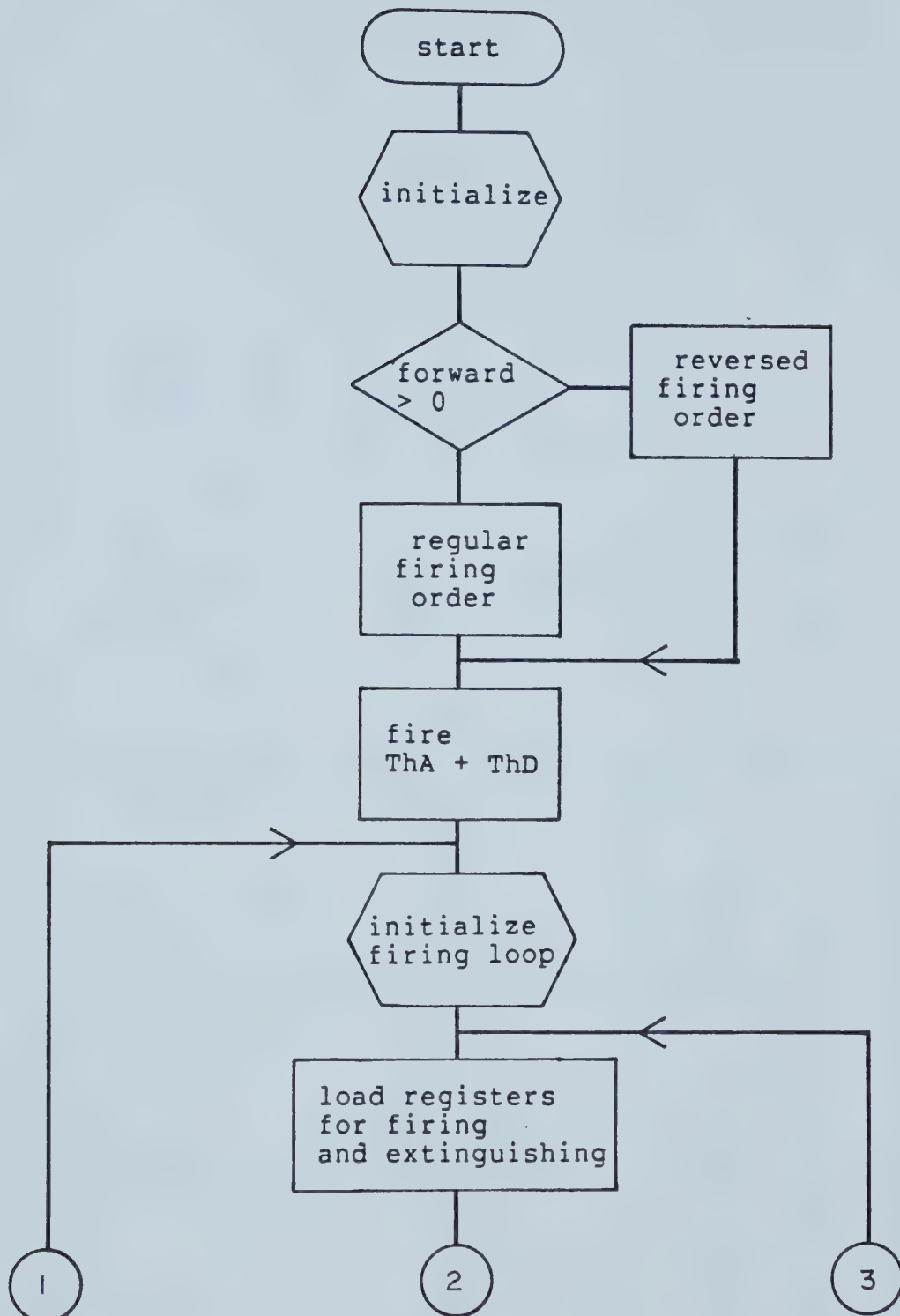


Figure 5.2 Subroutine "start"

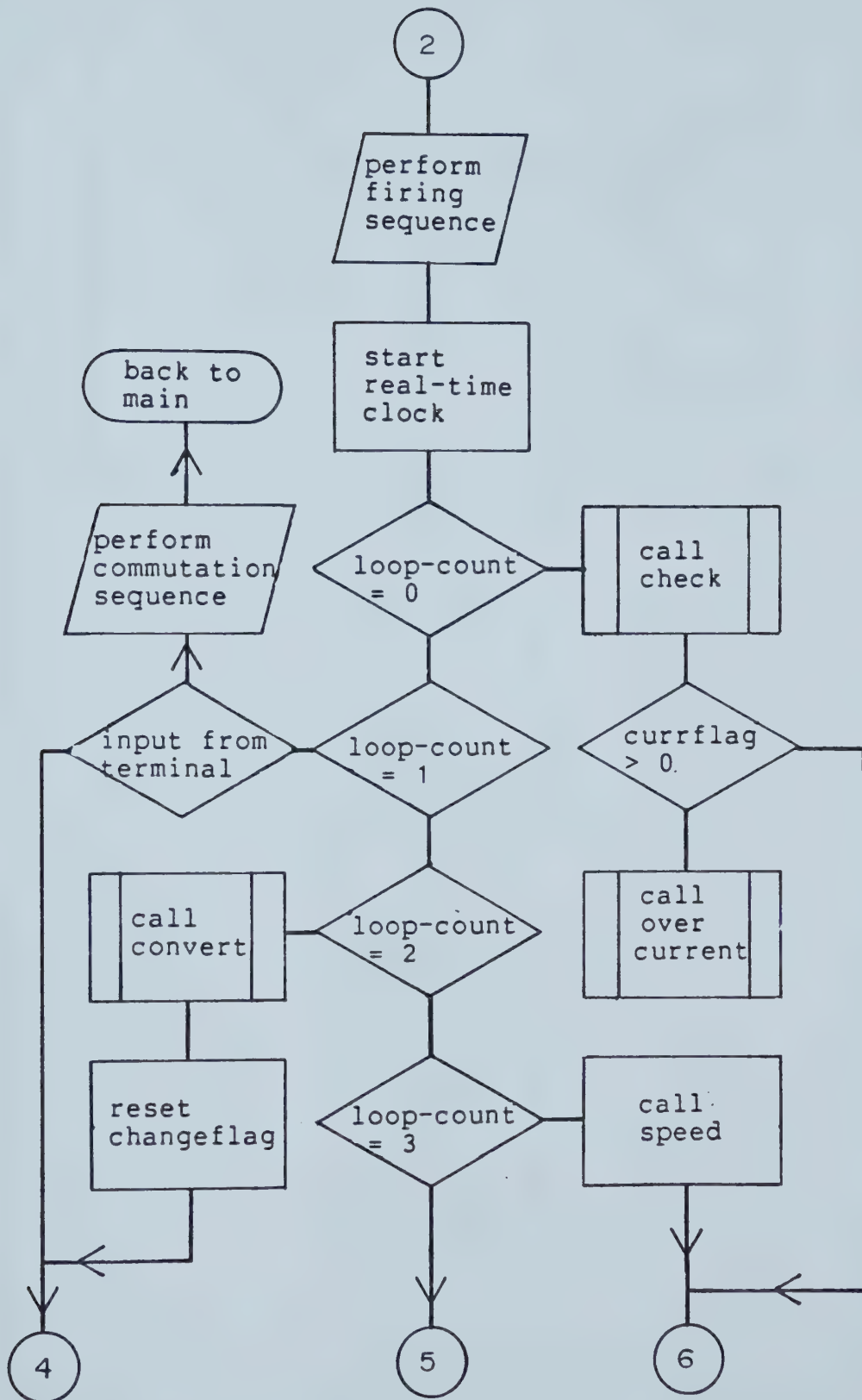


Figure 5.2 continued

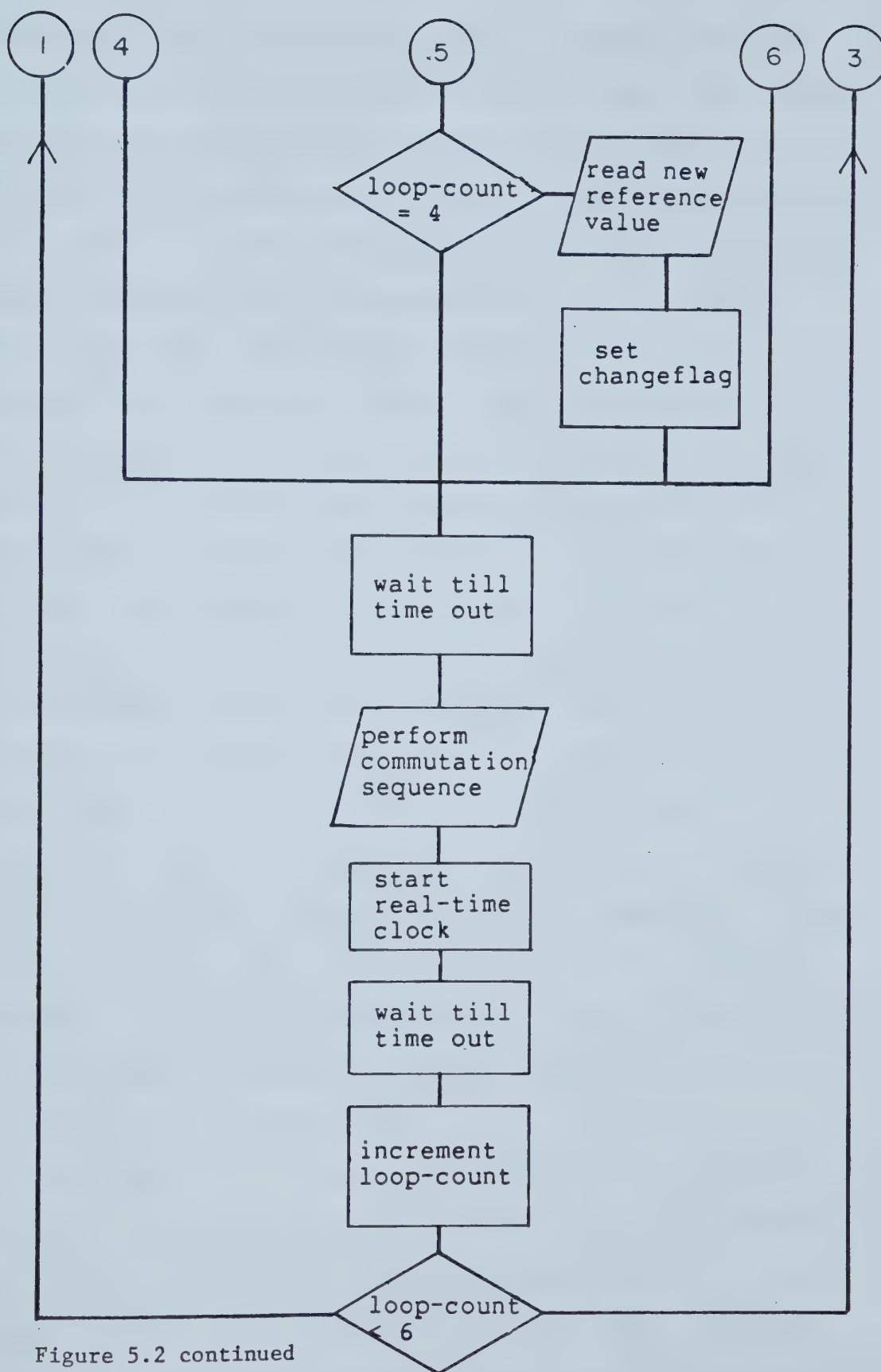


Figure 5.2 continued

completing a full cycle of the output voltage. After the firing pulse has been put out to the inverter, time remains for performing other tasks. Depending on the desired stator frequency, the remaining time is from 2.5ms (for frequencies up to 60 Hz) to 1.5 ms (for maximum frequency). A check of system variables is only necessary every full cycle of output frequency, hence the particular task is only performed every sixth run through the loop. Depending on the value of a loop counter and after checking certain flags, a subroutine is entered. After return from the routine a waiting loop is entered and nothing is done until the real-time clock signals overflow. When this signal is acknowledged by the CPU the end of the conduction interval is reached and a commutation thyristor pair is to be fired. The end of the nonconduction interval α is detected in another loop. This time can not be used for execution of subroutines, since at frequencies above 30 Hz the remaining time would be insufficient. The order of commutation pulses is not arbitrary, since only one half of the bridge is commutated at a time as described in Chapter 3. The constants EGON and FRIDA are for extinguishing the positive and negative half respectively.

The "start" routine also catches interrupt signals that might occur. For instance the "convert" routine signals a floating point exception when a very small value for the frequency has been read and divide by zero has occurred. Detection of interrupt signals is essential for ensuring

safe shut down of the inverter. If the signals were not detected, the program would be terminated immediately leaving, the main thyristors in the conducting state, which would destroy the inverter. In order to safely shut it down, the routine "interrupt" is called, which fires the appropriate commutation thyristors, then writes a message to the terminal and terminates execution of the program.

5.3 Routine "convert"

As was pointed out earlier, the control strategy is slip control but underlying this is a voltage control mechanism. The average value of the output voltage is varied in three intervals. From 0 to 30 Hz the output voltage as a function of frequency has a different slope than from 30 to 60 Hz. This is necessary for compensating the increased effect of the winding resistances. The average value of the output voltage as a function of frequency is shown in Figure 5.3. In the interval from 30 to 60 Hz, the output voltage rises linearly to a maximum value. This is accomplished by keeping the conduction interval β constant and varying the nonconduction interval α . The line to line voltage for the pulse width modulation inverter has a six step shape which is shown in Figure 5.3.

The average output voltage can be computed from Fourier-analysis,

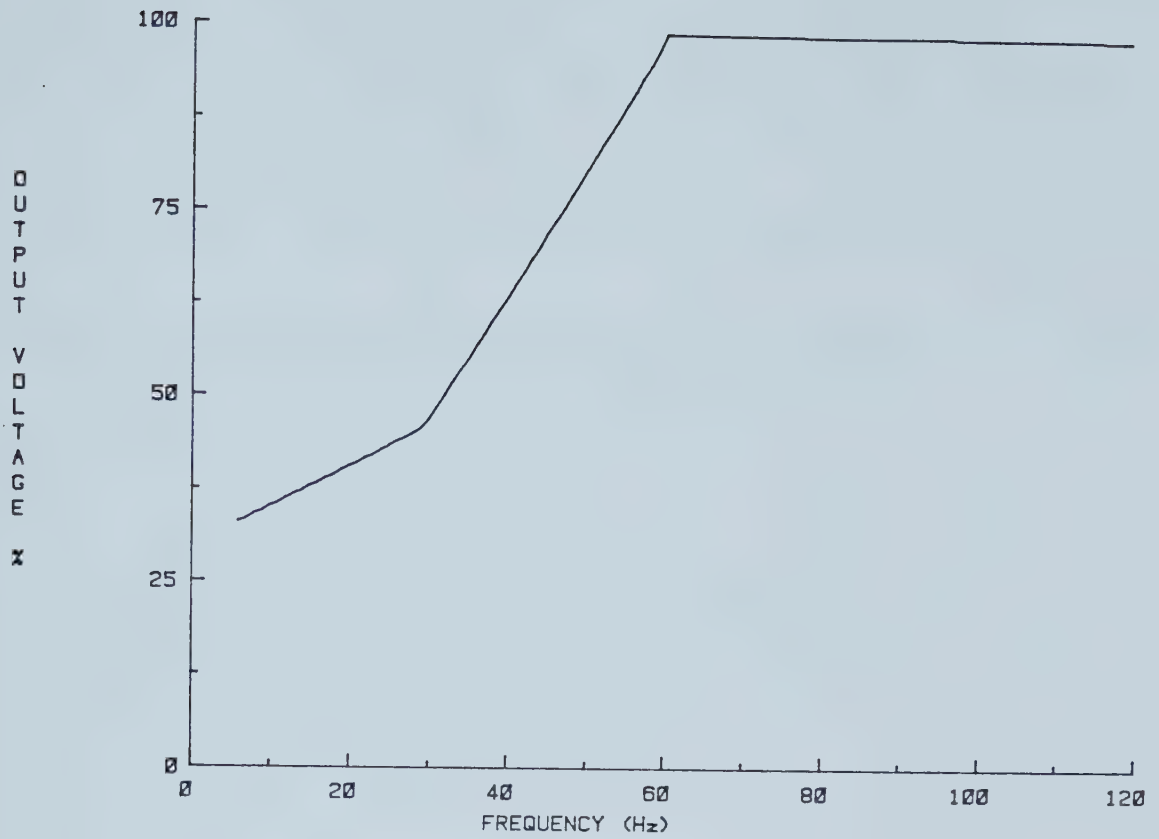


Figure 5.3 Output voltage versus frequency

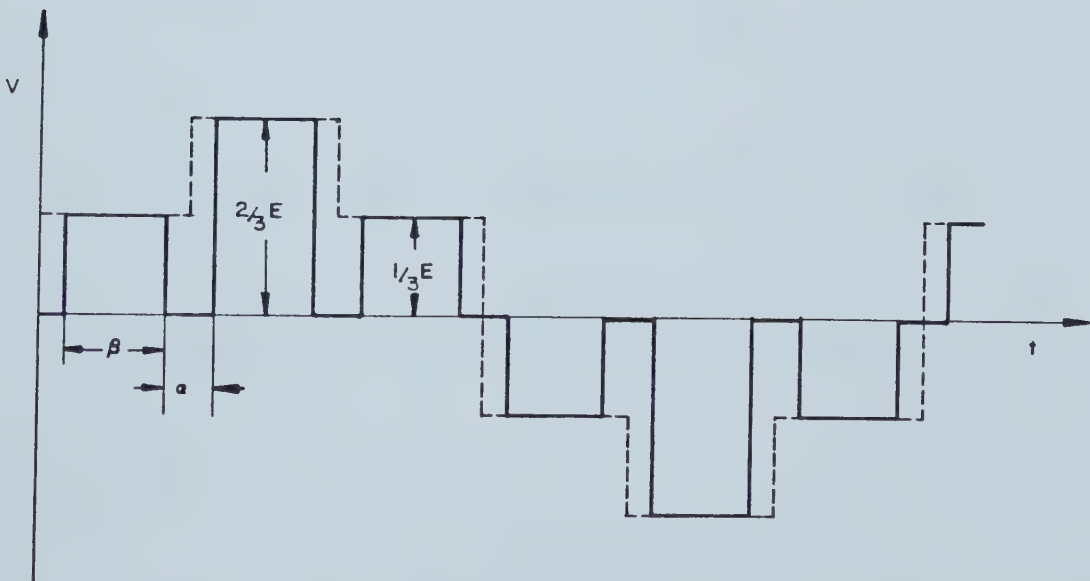


Figure 5.4 Inverter output voltage

$$V_{\text{out}} = \sum_{n=1}^{\infty} b_n \sin n\omega t \quad (5.1)$$

with

$$\begin{aligned} b_n &= \frac{1}{\pi} \int_0^{\pi} V(t) \sin n\omega t \, d\omega t = \frac{4}{\pi} \int_0^{\frac{\pi}{2}} V(t) \sin n\omega t \, d\omega t \\ &= \frac{4}{\pi} \int_{\frac{\pi}{6} - \frac{\beta}{2}}^{\frac{\pi}{6} + \frac{\beta}{2}} \frac{1}{3} E \sin n\omega t \, d\omega t = \frac{4}{\pi} \int_{\frac{\pi}{2} - \frac{\beta}{2}}^{\frac{\pi}{2}} \frac{2}{3} E \sin n\omega t \, d\omega t \end{aligned} \quad (5.2)$$

$$= \frac{4 E}{3n\pi} \left[(-\cos n\omega t) \Big|_{\frac{\pi}{6} - \frac{\beta}{2}}^{\frac{\pi}{6} + \frac{\beta}{2}} + 2(-\cos n\omega t) \Big|_{\frac{\pi}{2} - \frac{\beta}{2}}^{\frac{\pi}{2}} \right] \quad (5.3)$$

The first term of Equation 5.3 cancels, so it simplifies to,

$$b_n = \frac{8 E}{3n\pi} \left(\cos n\frac{\pi}{2} \cos n\frac{\beta}{2} + \sin n\frac{\pi}{2} \sin n\frac{\beta}{2} - \cos n\frac{\pi}{2} \right) \quad (5.4)$$

which is,

$$b_n = \frac{8 E}{3n\pi} (-1)^n \sin n\frac{\beta}{2} \quad n=1,3,5\dots \quad (5.5)$$

and,

$$b_n = \frac{8 E}{3n\pi} (-1)^n (1 - \cos n\frac{\beta}{2}) \quad n=2,4,6\dots \quad (5.6)$$

A more general expression can be found with different indices,

$$b_n = \frac{8 E}{3n\pi} \left((-1)^k \sin(2k+1)\frac{\beta}{2} + (-1)^i (1 - \cos 2i\frac{\beta}{2}) \right) \quad (5.7)$$

where

$$2k+1 = n = 2i \quad n = 1, 2, 3, 4, \dots$$

The fundamental is given by,

$$b_1 = \frac{8 E}{3\pi} \sin \frac{\beta}{2} \quad (5.8)$$

and V_{out} is given by,

$$V_{out} = \frac{8 E}{3\pi} \sin \frac{\beta}{2} \sin \omega t \quad (5.9)$$

The maximum value for beta, however, is $\frac{\pi}{3}$ and so the argument of the sin will be $\frac{\pi}{6}$ at the maximum. For these small values the sin may be approximated by the argument and the relationship between the output voltage and the conduction interval is now linearized. Hence it follows that,

$$V_{out} = \frac{8 E}{3\pi} \frac{\beta}{2} \sin \omega t = k \frac{\beta}{2} \sin \omega t \quad (5.10)$$

or, with the frequency f , given by,

$$f_1 = \frac{1}{6(\alpha + \beta)} \quad (5.11)$$

so that,

$$\frac{V_{out}}{f_1} = \frac{8 E}{3n\pi} 6(\alpha + \beta) \sin \omega t \quad (5.12)$$

It can be seen that keeping the interval beta constant while varying alpha, will vary the frequency but keep the ratio V/f constant.

For frequencies in excess of 60 Hz, the average voltage remains constant as the frequency increases. This is equivalent to decreasing the flux and has the same effect as field-weakening for a dc-shunt motor. These considerations have been implemented in the routine "convert", which takes the frequency as an argument and, depending on the range, performs the necessary calculations for the intervals alpha and beta. The real-time-clock of the SNORKEL can only be programmed in integer values and consequently the reciprocal of the frequency has to be converted to an integer without increasing the error considerably. This is done by multiplying the reciprocal of the frequency by 10^6 which gives the period in μs . The necessary computations for determining the intervals alpha and beta are performed taking into account the system delay time. This is approximately 200 μs and is the time it takes the microcomputer to come out of the waiting loop and reinitialize the timing loop in the routine start. The routine returns to the calling function after setting pointers to the new addresses for alpha and beta.

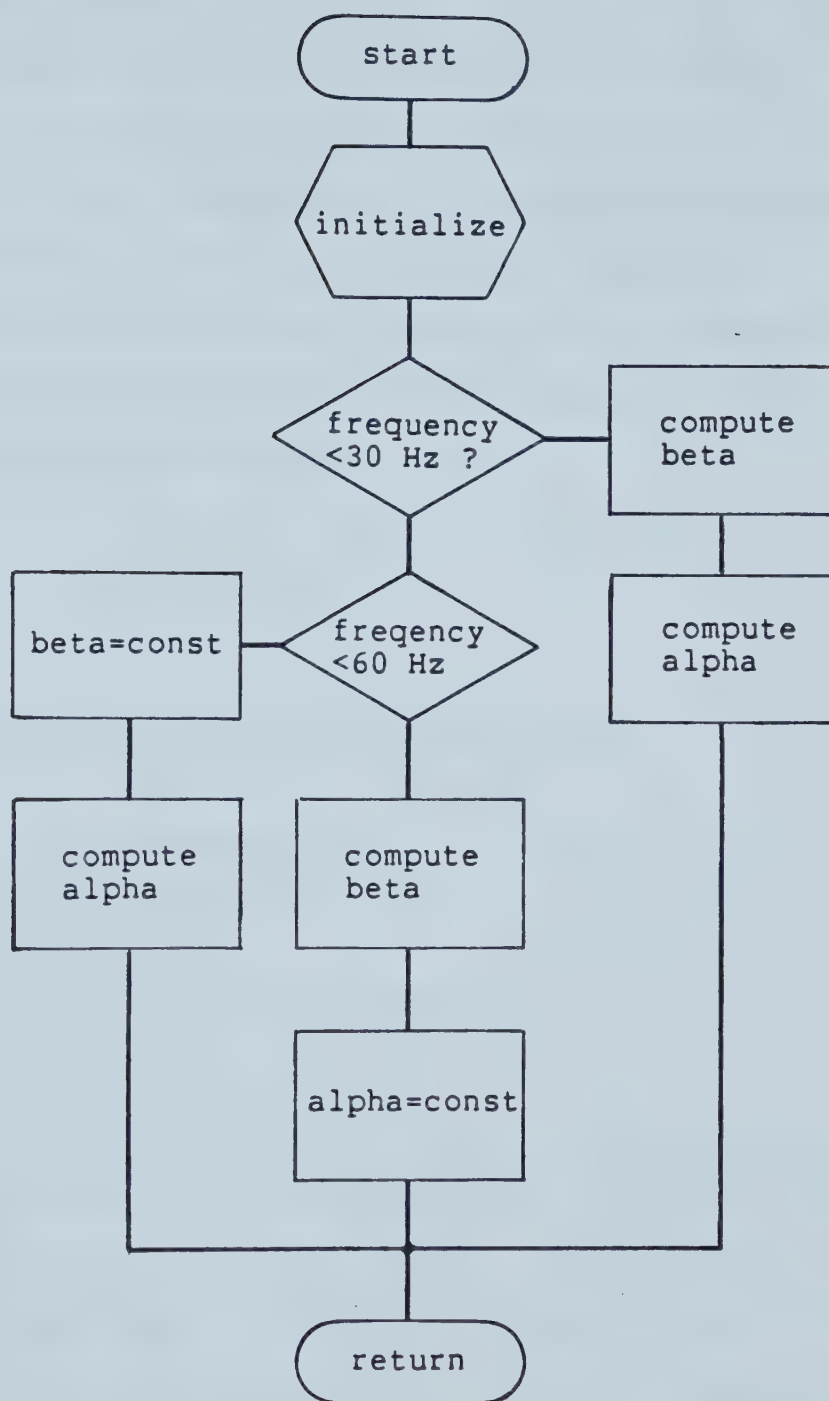


Figure 5.5 Flowchart for the routine "convert"

5.4 Routine "speed"

The actual speed control is performed in this subroutine using a digital proportional-integral- derivative control algorithm (PID control). A schematic of an analog PID controller is shown in Figure 5.6. The three sub-controllers each compute a change of the output signal from the error signal and the three outputs are added together and passed to the process.

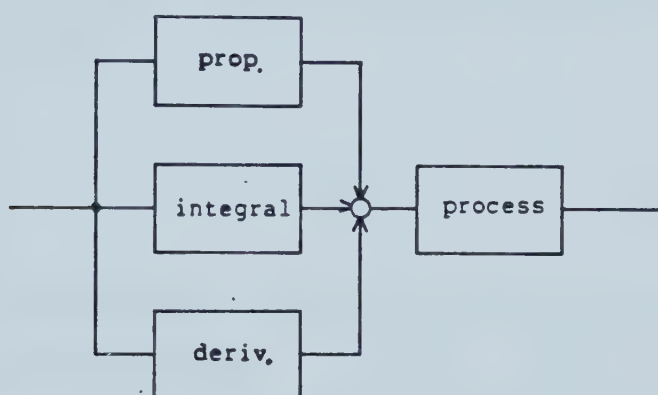


Figure 5.6 Schematic of PID controller

In the digital controller, output signals are computed from the differences of measured signals at times $t, t-1, t-2$. For the induction motor slip control, the algorithm works on a signal proportional to the rotor frequency and a reference setting for slip. The rotor shaft frequency is measured by a magnetic pickup that senses the different reluctance between teeth and gaps of a gear attached to the rotor. The gear was manufactured to give 100 pulses per revolution in order to keep the pickup error down to 1%. After some pulse-shaping,

this signal was fed into a frequency to voltage converter and then passed to the microcomputer analog channel. This method of signal processing was chosen in order to keep the interfacing hardware to a minimum and to make use of the facilities of the SNORKEL. It is clear that because of the double conversion a higher error can be expected. In a subsequent shaft speed detector system a direct digital conversion should be used.

The routine "speed" computes the new stator frequency from Equation 5.13

$$f_1 = f_1(n-1) + \Delta \quad (5.13)$$

The error signal Δ is composed of the integral, derivative and proportional parts, which are computed by the following difference equations.

The proportional action is given by,

$$\begin{aligned} \Delta p &= K_p (e(n) - e(n-1)) \\ &= K_p (r(n) - f_2(n) - r(n-1) - f_2(n-1)) \end{aligned} \quad (5.14)$$

Since the reference signal is assumed to be unchanged this leads to,

$$\Delta p = K_p (f_2(n-1) - f_2(n)) \quad (5.15)$$

Integral action is described by the instantaneous equation,

$$\Delta i = K_i \int e \, dt \quad (5.16)$$

The discrete form of this equation is,

$$\frac{\Delta i}{\Delta t} = K_i e(n) = K_i (r(n) - f_2(n)) \quad (5.17)$$

and hence,

$$\Delta i = K_i \Delta t (r(n) - f_2(n)) = K_i T (r(n) - f_2(n)) \quad (5.18)$$

with T =sampling time

Derivative action is accomplished with the following algorithm,

$$\begin{aligned} \Delta d &= K_d \frac{de(n)}{dt} - K_d \frac{de(n-1)}{dt} \\ &= K_d \left[\frac{e(n)-e(n-1)}{\Delta t} - \frac{e(n-1) - e(n-2)}{\Delta t} \right] \end{aligned} \quad (5.19)$$

$$= \frac{K_d}{T} (r(n)-f_2(n)-r(n-1)+f_2(n-1)-r(n-1)+f_2(n-2)+r(n-2)) \quad (5.20)$$

$$= \frac{K_d}{T} (2f_2(n-1) - f_2(n-2) + f_2(n)) \quad (5.21)$$

The output signal is formed by adding all the components together, yielding,

$$\begin{aligned}
 f_1(n) &= f_1(n-1) + \Delta p + \Delta i + \Delta d \\
 &= f_1(n-1) + K_p (f_2(n-1) - f_2(n)) \\
 &\quad + K_i T (r(n) - f_2(n)) \\
 &\quad + \frac{K}{T} d (2f_2(n-1) - f_2(n-2) - f_2(n))
 \end{aligned}
 \tag{5.22}$$

The new stator frequency is then computed according to Equation (5.13) and control is passed to the "start"-routine again.

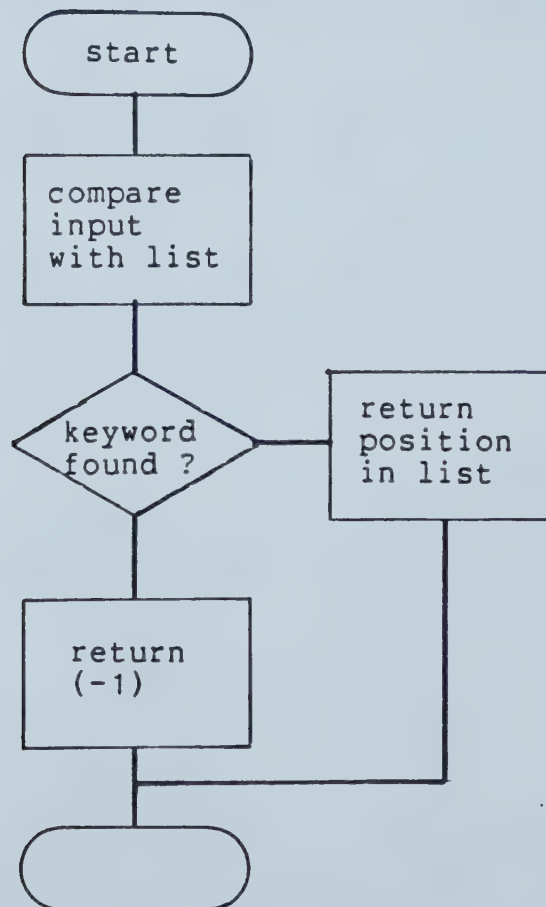


Figure 5.7 Flowchart for routine "compare"

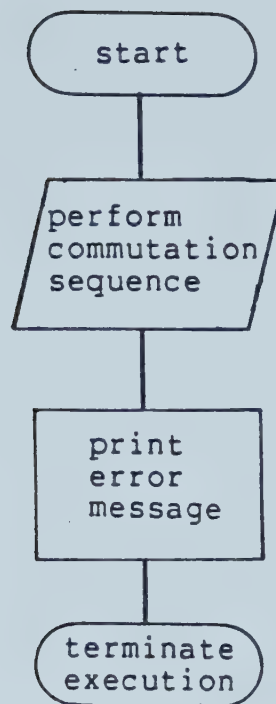


Figure 5.8 Flowchart for the routines "interrupt" and "overcurrent"

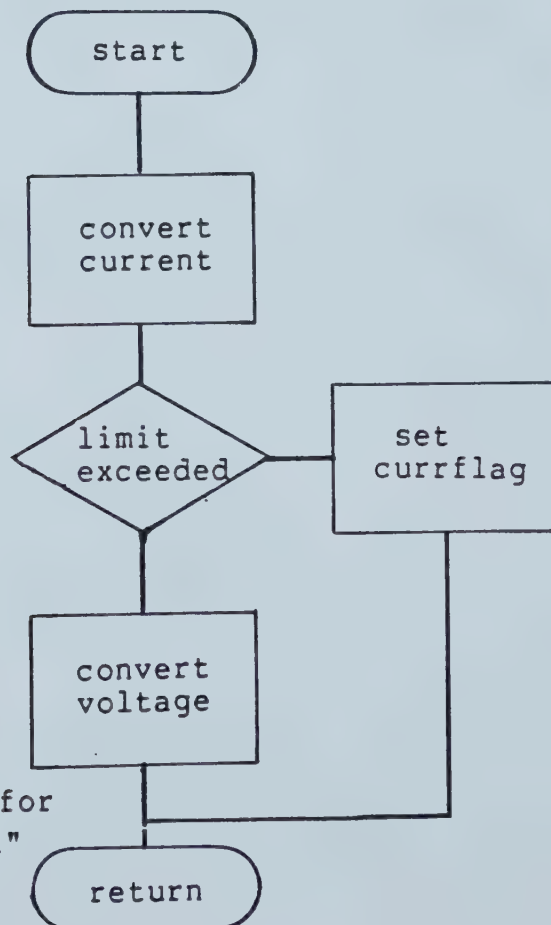


Figure 5.9 Flowchart for routine "checkcurrent"

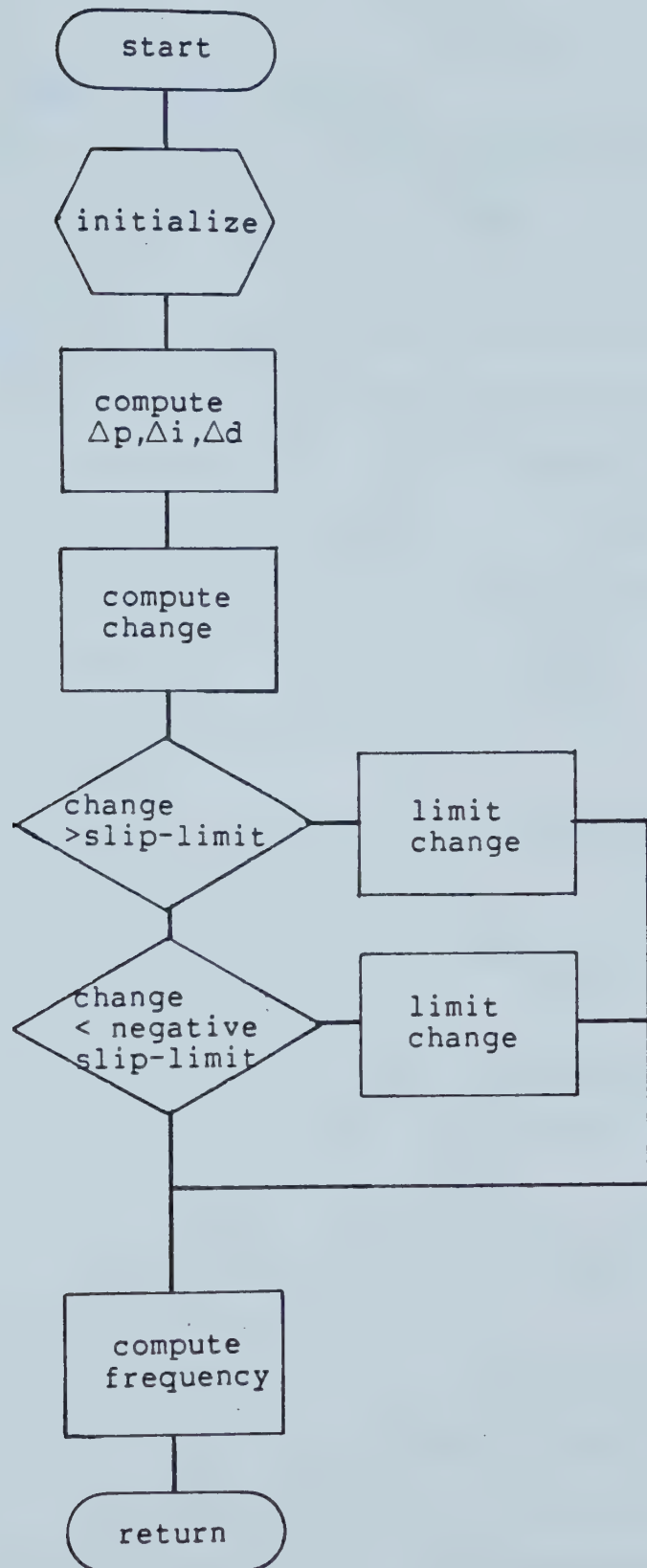


Figure 5.10 Flowchart for the routine "speed"

6. Experimental Results

The equipment described so far was connected together as shown in Figure 6.1. The inverter was fed through a autotransformer which served as a variable ac-source for the inverter. The best results were obtained when adjusting the voltage so that a dc-bus voltage of 150 V was measured with the machine not running. With this magnitude of voltage the motor can be started at any desired frequency. The dc-bus voltage drops, however, when the motor is started and during normal operation. This voltage drop is due to the large filter inductor. The magnitude of this choke could be reduced in a subsequent design.

The inverter proved to be sufficiently fast to produce a pulse width modulated output waveform. However, as mentioned in Chapter 3.5, the design of the commutation circuit is suboptimal and therefore performance of the circuit can be improved by recalculating the components in the commutation circuit. The values of the commutation elements L_1 , L_2 , and C , in the current design give a commutation time t of 120 μ s. This is the time it takes for the inverter bus voltage to rise from $-E$ to $+E$ at commutation instant. This time can be safely reduced to 40 μ s. The size of capacitor C , should be increased since this improves the current carrying capability of the inverter. With increasing capacitance the inductors will be decreased accordingly. The value can be determined with the formulae,

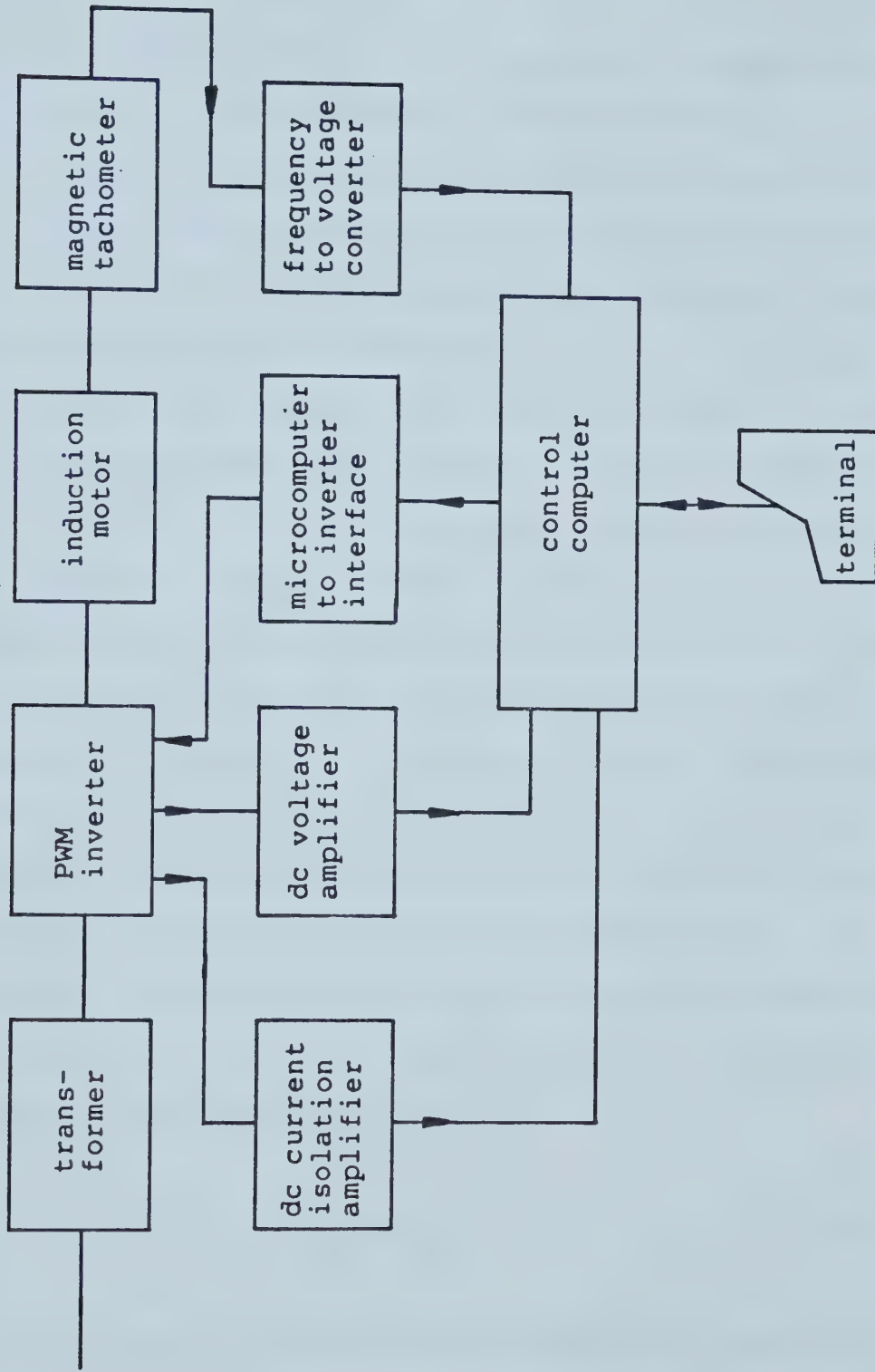


Figure 6.1 Experimental configuration for testing the PWM-inverter

$$\omega_0 = \frac{1}{\sqrt{L C}} = \frac{2\pi}{T} \quad (6.1) \quad \text{and,} \quad t_c = \frac{1}{2} T = \frac{\sqrt{L C}}{4 \pi} \quad (6.2)$$

since t is approximately $1/2$ of the period. Suggested values for C , and L , are $5.0\mu\text{F}$ and 1mH respectively.

A further improvement can be expected by using another type of diode for the freewheeling diodes DA and DB. Using devices with more current handling capability will render the resistors RA and RB unnecessary and will therefore reduce commutation losses. This is of particular interest at low frequencies since then losses in the two resistors are high. The current in the freewheeling path has the same magnitude at all frequencies but the duration of its flow through the path is longer at frequencies from 10 to 50 Hz. At frequencies higher than 50 Hz the commutation is almost immediately followed by a conduction interval. When the main thyristors are gated the current that was flowing in the freewheeling path is diverted into the motor and therefore the losses in resistors RA and RB are reduced. At considerably high frequencies the commutation losses rise again since the number of switching cycles increases and the dissipated power is proportional to

$$P_{\text{diss}} \propto W_{\text{comm}} f \quad (6.3)$$

A more sophisticated way of reducing the commutation losses is to use a three-winding pulse transformer as suggested in reference [8]. Rather than have the commutation

energy dissipated in resistors it can be fed back into the filter capacitor and serve to improve the overall efficiency of the inverter.

The control computer proved to be sufficiently fast to drive the inverter in the frequency range from 10 to 100 Hz. As mentioned in Chapter 5.3, the output voltage of the inverter has a six step shape with each step consisting of a conduction interval β and a nonconduction interval α . Between the firing pulse for a particular conduction interval and the extinguishing pulse, time remains for the computer to perform background operations, like measuring current and voltage and performing the speed control task. The interval α is varied in wide limits and therefore can not be used for background operations. The minimum conduction interval is 1.4 ms (this is at 100 Hz) and at this point the microcomputer was capable of performing all the required functions. The a/d conversion routine "check" which measures two quantities requires a large amount of time and is a crucial point for the limitation of output frequency.

SNORKEL is not fast enough for switching more than six times per full cycle of output voltage (at least with the program as it is now). If a multiply switched output waveform is desired the inverter part should be done with transistors, which would eliminate commutation circuits, reduce losses, and would facilitate very short conduction intervals.

6.0.1 Measurements

When making measurements on the inverter circuit one should be aware of the fact that metallic parts of the inverter can be connected with the mains and that the positive and negative dc-bus potential, when grounded through the oscilloscopes ground lead, will be short-circuited. The reason for this is that no isolation transformer is used to provide the variable ac and the autotransformer always gives a direct connection to the mains.

Hence it is important that the oscilloscope be operated through an isolation transformer. Furthermore one should assure that no conducting part of the casing of the oscilloscope is in touch with conducting parts of other equipment. If one of these two conditions is not satisfied, shortcircuits can occur and semiconductors will be destroyed.

The currents can be measured using the shunts that were included in the inverter design. It is less failure-prone, however, and better results can be obtained, using an HP 456A ac current probe that can be clipped around any insulated conductor. The probe gives an output voltage of 1 V/A.

The picture shown in Figure 6.2 shows the current of phase A on the upper trace and the voltage V on the lower trace, taken at a frequency of 60 Hz. Figure 6.3 shows the same measurement values but at a frequency of 15 Hz. The

arrow is pointing to the interval in which the feedback diodes are conducting. The voltage seen in this interval is the motor-e.m.f.

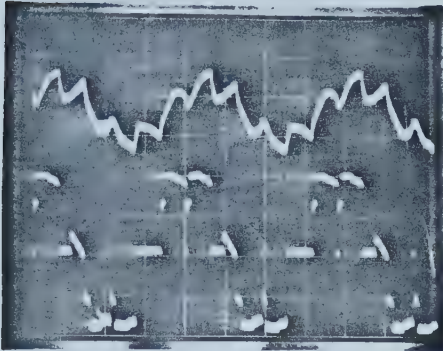


Figure 6.2 Phase current and line to line voltage at 60 Hz. Sensitivity : 2A/div, 50V/div time : 5ms/div

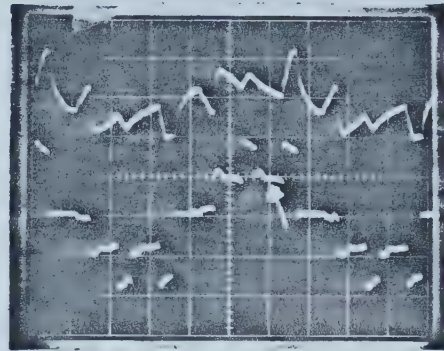


Figure 6.3 Phase current and line to line voltage at 15 Hz. Sensitivity : 2A/div, 50V/div time : 10ms/div

The blocking voltage of the commutation thyristor Th_C and the current through this device are shown in Figure 6.4. The waveforms shown in Figure 6.5 through 6.7 are pictures of the commutation current and the dc-bus voltage (lower trace). From these figures it can be seen that the voltage has an appreciable overshoot caused by the resistors R_A or R_B respectively. Figure 6.6, taken at a frequency of 60 Hz, shows that a triplet of thyristors is fired immediately after completion of the commutation. The duration of the voltage overshoot is reduced since the current through the feedback path is now diverted into the main thyristors. The last two pictures show the firing pulses for thyristor Th ,

(upper trace) and ThB+ThC (lower trace), taken at different output frequencies.

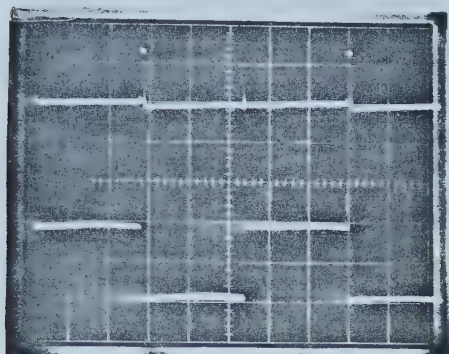


Figure 6.4 Commutation current and blocking voltage.
Sensitivity: 2A/div, 50V/div
time: 1ms/div

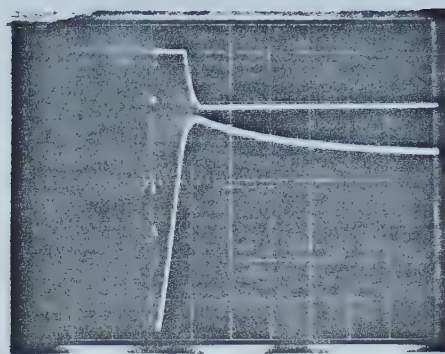


Figure 6.5 Commutation current and dc-bus voltage.
Sensitivity: 2A/div, 50V/div
time: 0.2ms/div

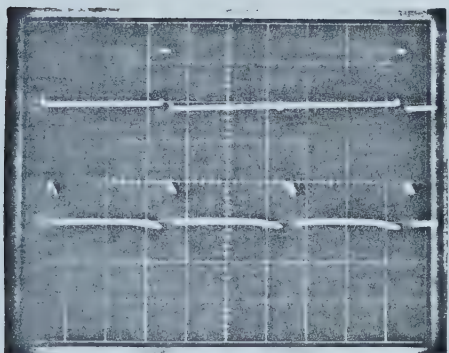


Figure 6.6 Commutation current and dc-bus voltage.
@60 Hz
Sensitivity: 2A/div, 50V/div
time: 1ms/div



Figure 6.7 Commutation current and dc-bus voltage.
@15 Hz
Sensitivity: 2A/div, 50V/div
time: 10ms/div

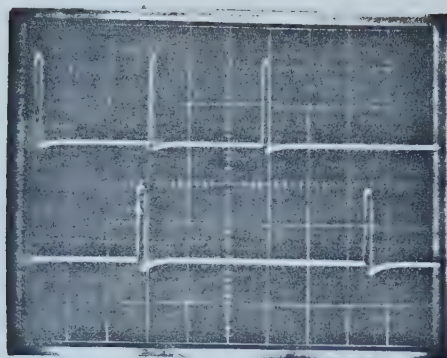


Figure 6.8 Firing pulses Th,
and Th_B+ Th_C. @60 Hz
Sensitivity: 2V/div, 2V/div
time: 1ms/div

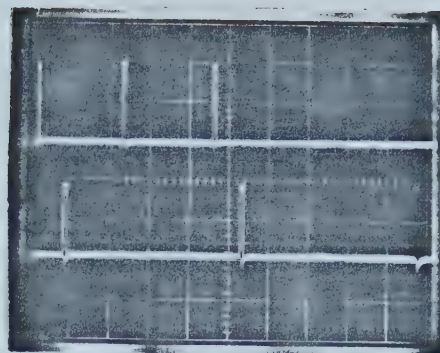


Figure 6.8 Firing pulses Th,
and Th_B+ Th_C. @15 Hz
Sensitivity: 2V/div, 2V/div
time: 5ms/div

Figure 6.10 shows recording of the motor speed for a manual step change in reference signal. Future projects could involve programming the microcomputer for different starting sequences using selectable reference value profiles.

Figure 6.11 shows a transient for the motor speed where the motor speed is controlled by the "softstart" command of the control program.

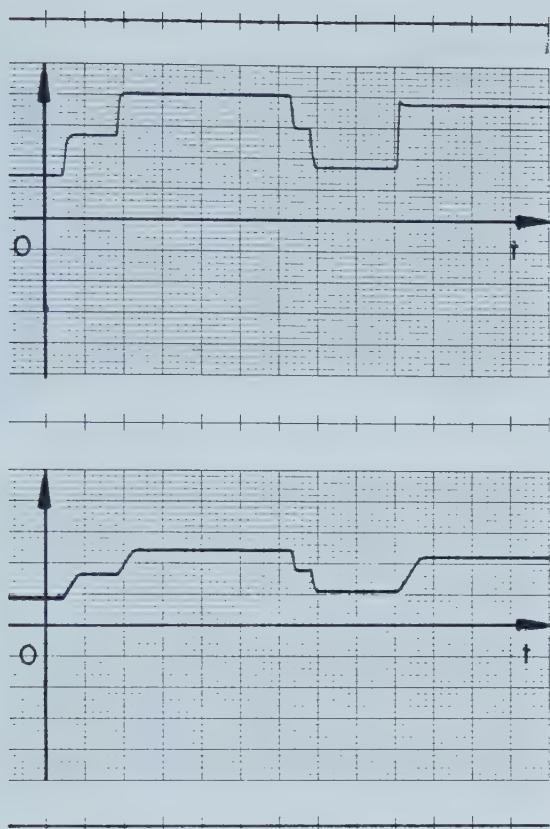


Figure 6.10 Stepresponse of rotor speed. Upper trace: reference, 120 rpm/div
Lower: speed, 100 rpm/div
time: 1 sec/div

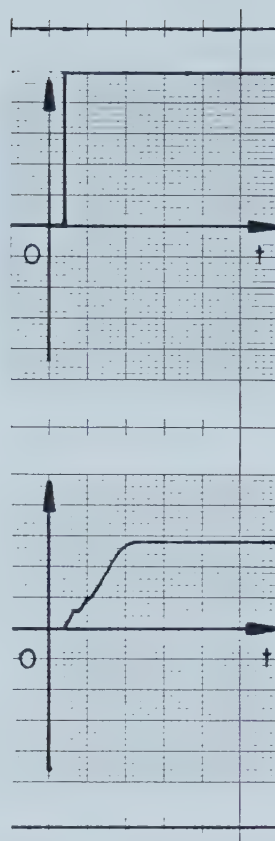


Figure 6.11 Rotor speed vs. time for softstart. Upper: reference, 120 rpm/div
Lower: speed, 100 rpm/div
time: 1 sec/div

The response to changes in motor load conditions is shown in Figure 6.12 and 6.13. Figure 6.12 shows the dc-generator current and rotor speed for open-loop conditions (no shaft-speed signal feedback to the microcomputer). Figure 6.13 illustrates transient recordings of the same signal for the closed loop-case with proportional control. With this latter controller, a change in load current of 20% results in a steady state error of about 8.4%.

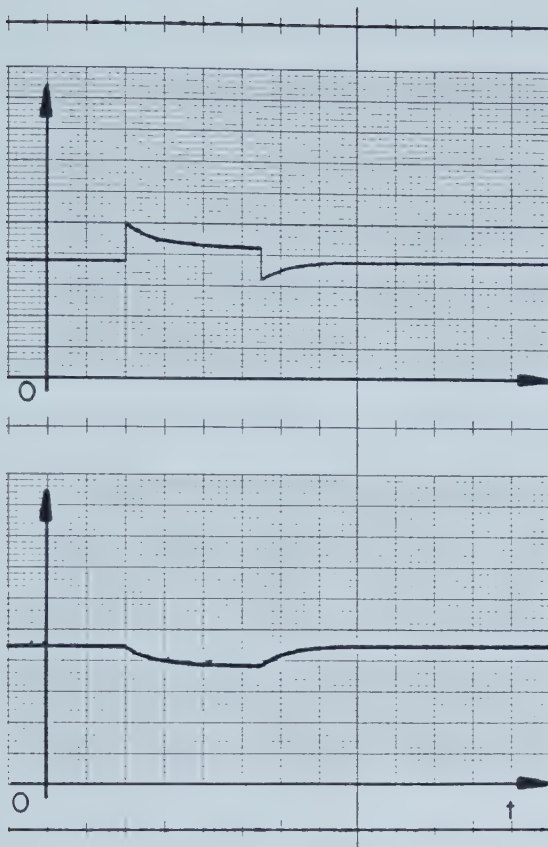


Figure 6.12 Stepresponse for load change. Upper:
Load current, 200 ma/div
Lower: speed, 60 rpm/div
time: 1 sec/div

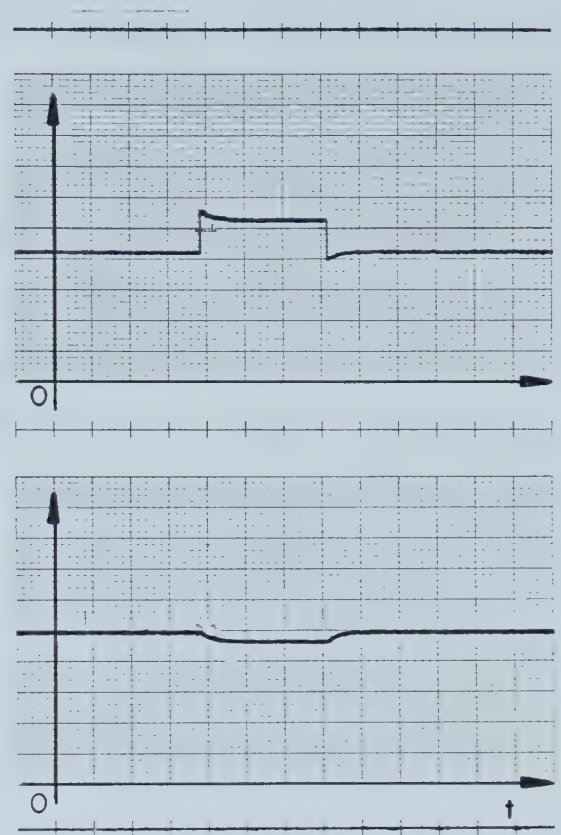


Figure 6.13 Stepresponse with feedback. Upper: Load current, 200 ma/div. Lower: rotor speed, 60 rpm/div.
time: 1 sec/div

7. Conclusion

This thesis presents theoretical considerations and hardware design description for a speed controller of an induction motor using a microcomputer. The supply for the motor consists of a three-phase, pulse-width modulated voltage, which is synthesized by the microcomputer gating the thyristors in a dc-side commutated inverter. The possible applications for a variable speed drive using an induction motor are pointed out. Applications include drives for large fans, centrifugal pumps, conveyor belts or similar tasks where a dc-motor cannot be applied because of adverse environmental conditions. The use of a microcomputer as controller for the inverter and the advantages of its use are described.

An introduction to the fundamentals of variable ac-drives is given, including a description characteristics of induction motors operated from a nonsinusoidal supply. Various schemes for generating such a supply and several choices for control strategies are stated. It is pointed out that the most reasonable method of controlling the induction motor is the constant slip control and this strategy was employed in the study.

The inverter used in this study as well as the characteristic features of the motor and the control computer are described, giving detailed design considerations.

A description is given of the software for generating the gate pulses for the inverter and the control routines, a list of suggestions for improving the inverter performance and a chapter with results of the design.

It is hoped that the inverter and the software package developed in this project will be a valuable tool for further research on variable speed-ac drives in the Department of Electrical Engineering at the University of Alberta.

8. Bibliography

General literature

- 1 Siemens "System Based Drive Technology"
Publication of Siemens AG, Erlangen 1979
- 2 J.M.D. Murphy "Thyristor Control of AC Motors"
Pergamon Press, Oxford, 1973
- 3 T.H.Barton "IEEE Short Course on Variable Speed Electric Drives"
IEEE-Publication, New York, 1981
- 4 Robert Joetten "Leistungselektronik Band1: Stromrichter Schaltungstechnik"
Robert Joetten, Vieweg, Braunschweig, 1977
- 5 Philip L. Alger "Induction Machines"
Gordon and Breach, New York, 1970
- 6 R.G.Bedfort and R.G.Hoft "Principles of Inverter Circuits"
R.G.Bedfort John Wiley and Sons, New York, 1964
- 7 General Electric Co. "Power Converter Handbook"
Canadian General Electric Co. 1976
- 8 John Harnden, Forest B. Golden (Editors) "Power Semiconductor Applications Volume I+II "
IEEE Press, New York, 1972
- 9 "Electrical Variable Speed Drives"
Conference Publication Nr.93, IEE, London, 1972
(U.K.)
- 10 "Electrical Variable Speed Drives"
Conference Publication Nr.179, IEE, London, 1979
(U.K.)

Inverters and Variable Speed Drives:

- 11 Wilhelm Leitgeb "Zur Bemessung drehzahlveraenderlicher Antriebe konstanter Leistung mit stromrichtergespeisten Drehfeldmaschinen"
ETZ-A, Berlin, 1973, Bd.94
- 12 J.Lindsay "Measurement Problems In Determining The

Efficiency Of Thyristor-Supplied Motor Drives
 Industry Applications Society Annual General
 Meeting 1977
 IEEE Press New York, 1977

- 13 M.Abbas,D.W.Novotny "Stator Referred Equivalent
 Circuits For Inverter Driven Electric Machines"
 Industry Applications Society Annual General
 Meeting 1978
 IEEE Press New York, 1978
- 14 D.M.Divan,T.H.Barton "Commutation Circuit Optimization
 For The Mc Murray Inverter"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 15 K.E.Dahlberg,M.Glogolja "Advancing The State Of The Art
 in Variable Speed Controls"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 16 T. Hori, M. Koido N. Matsudaira, T. Miyata "Shockless
 Start Of Large Capacity Squirrel-Cage Induction
 Motors For Driving Pumps Or Fans"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 17 B.De Fornel,J.M.Farines,J.C.Hapiot "Numerical
 Estimation Of The Speed Of An Asynchronous Machine
 Supplied By A Static Converter"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 18 Y.Dote,K.Anbo "Combined Parameter And State Estimation
 Of Controlled Current Induction Motor Drive System
 Via Stochastic Linear Filtering Technique"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 19 L.J.Garces "Parameter Adaption For The Speed Controlled
 Static AC Drive With Squirrel-Cage Induction Motor"
 Industry Applications Society Annual General
 Meeting 1979
 IEEE Press New York, 1979
- 20 B.Kampschulte "Comparison Of An Asynchronous Machine
 Drive With Current-Source Inverter And
 Voltage-Source Inverter"
 Industry Applications Society Annual General

Meeting 1979
IEEE Press New York, 1979

- 21 R.Wloydka "Power Characteristic Comparisons Of AC Traction Inverter Systems"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 22 F.ShibataA.Ohtsubo,KTsuruta,T.Kohrin "Speed Control Of A Cascade Induction Motor With Three Sets Of Converters In Its Secondary Circuit"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979

Microprocessor applications to motor control:

- 23 H.Le-Huy,R.Perret,D.Roye. "Microprocessor Control of a Current-Fed Synchronous Motor Drive"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 24 H.Le-Huy,R.Perret,D.Roye. "A Digital Firing Scheme for Thyristor Converters"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 25 J.Peacock,T.R.Visnawathan, R.S.Ramshaw "A Microprocessor Based System For Testing And Controlling Induction Motors"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 26 R.Gabriel,W.Leonhard,C.Nordby "Field Oriented Control Of Standard AC-Motor Using Microprocessors"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 27 V.Athani,S.Deshpande "Microprocessor Control Of A Three Phase Inverter Used In Induction Motor Speed Control System"
Industrial Applications of Microprocessors
IECI 1980
- 28 F.Harashima,H.Hayashi "Dynamic Performance Of Current Source Inverter Fed Induction Motors"

Industry Applications Society Annual General
Meeting 1979

IEEE Press New York, 1979

- 29 A.R.Miles,D.Novotny "Transfer Functions Of The Slip Controlled Induction Machine"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 30 Dr.M.Ramamoorthy,M.Arunachalam "Dynamic Performance Of A Closed Loop Induction Motor Speed Control System With Phase Controlled SCRs In The Rotor"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 31 R.Gabriel,W.Leonhard,C.Nordby "Microprocessor Control Of Induction Motors Employing Field Coordinates"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 32 A.B.Plunkett,J.D.D'Atre,T.A.Lipo "Synchronous Control Of A Static AC Induction Motor Drive"
Industry Applications Society Annual General Meeting 1977
IEEE Press New York, 1977

Pulse Width Modulated Inverters:

- 33 E.Dwyer,B.T.Ooi "A 'Lookup Table'Based Microprocessor Controller For A Three Phase PWM Inverter"
Industrial Applications of Microprocessors
IECI 1979
- 34 J.M.D.Murphy,R.G.Hoft,L.S.Howard "Controlled Slip Operation Of An Induction Motor With Optimum PWM Waveforms"
Industrial Applications of Microprocessors
IECI 1980
- 35 G.Buja,P.Fiorini "A Microcomputer Based, Quasi-Continuous Output Controller For PWM Inverters"
Industrial Applications of Microprocessors
IECI 1980
- 36 P.Bhagwat,V.Stefanovic "Some New Aspects In The Design Of PWM Inverters"
Industry Applications Society Annual General

Meeting 1979
IEEE Press New York, 1979

- 37 S.B.Dewan,J.Caceres "An Input Current-Commutated Three-Phase PWM Inverter"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 38 I.Piel,S.Talukdar,P.Wood Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979"Characterization Of Programmed-Waveform PWM Modulation"
Industry Applications Society Annual General Meeting 1979
IEEE Press New York, 1979
- 39 H.Raphael "Additional Losses In PWM Inverter-Fed Squirrel Cage Motors"
Industry Applications Society Annual General Meeting 1977
IEEE Press New York, 1977
- 40 F.G.De Buck "Losses and Parasitic Torques In Electric Motors Subjected to PWM-Waveforms"
Industry Applications Society Annual General Meeting 1977
IEEE Press New York, 1977
- 41 T.L.Grant,T.H.Barton "A Highly Flexible Controller For A Pulse Width Modulation Inverter"
Industry Applications Society Annual General Meeting 1978
IEEE Press New York, 1978
- 42 A.Abbondanti "A Digital Modulator Circuit For PWM Inverters"
Industry Applications Society Annual General Meeting 1978
IEEE Press New York, 1978

LSI-11 Related Literature

- 43 "Snorkel"
"LSI-11 Software Development System Documentation"
Advanced Digital Engineering Corporation,
Saskatoon, 1981
- 44 "Microcomputer Processor Handbook"
digital corporation 1980

- 45 "Microcomputer Interfaces Handbook"
digital corporation 1980
- 46 "User Manual for DT2769-Realtime Clock"
Data Translation, Inc. 1979, Natick, Mass.
- 47 "LSI-11 Compatible 4 Channel D/A-Converter Subsystem"
Data Translation, Inc. 1978, Natick, Mass.
- 48 "LSI-11 Compatible Digital I/O-Board DT2768"
Data Translation, Inc. 1979, Natick, Mass.
- 49 Brian W. Kernighan and Dennis M. Ritchie "The C
Programming Language"
Prentice-Hall, Software Series, New York, 1978
- 50 "The UNIX Timesharing System" Volume I to IV, Seventh
Edition
Bell Telephone Laboratories, Murray Hill, N.J. 1979

Appendix

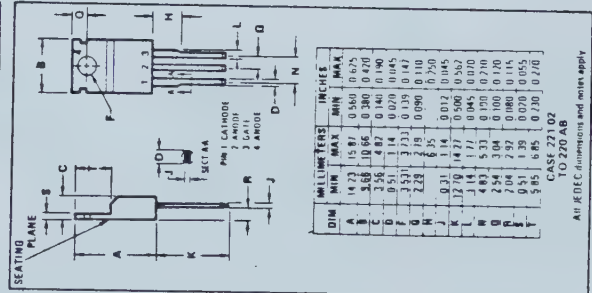
8.1 Semiconductor Data-Sheets



SILICON CONTROLLED RECTIFIERS

- designed primarily for half-wave ac control applications, such as motor controls, heating controls and power supplies, or where ever half wave silicon gate controlled, solid state devices are needed
- Glass Passivated Junctions and Center Gate Fire for Greater Parameter Uniformity and Stability
- Small, Rugged, Thermowatt Construction for Low Thermal Resistance, High Heat Dissipation and Durability
- Blocking Voltage to 800 Volts

THYRISTORS
12 AMPERES RMS
50-800 VOLTS



MAXIMUM RATINGS

Rating	Symbol	Value	Unit
Peak Reverse Blocking Voltage (1)	V_{RRM}	50	Volts
		100	
		200	
		400	
		800	
Forward Current RMS $T_J = 125^\circ\text{C}$	I_T	12	Amps
Peak Forward Surge Current (AI Conduction Angle)	I_{TSM}	100	Amps
Circuit Fusing Consideration ($T_J = 40$ to 125°C ; $t = 1.0$ to 8.3 ms)	I_2	40	A^2
Forward Peak Gate Power	P_{GM}	20	Watts
Forward Average Gate Power	P_{GAV}	0.5	Watt
Operating Junction Temperature Range	T_J	-40 to +125	$^\circ\text{C}$
Storage Temperature Range	T_{stg}	-40 to +150	$^\circ\text{C}$

THERMAL CHARACTERISTICS

Characteristic	Symbol	Max	Unit
Thermal Resistance, Junction to Case	$R_{\theta JC}$	2.0	$^\circ\text{C}/\text{W}$

(1) V_{RRM} for all types can be applied on a continuous dc basis without incurring damage. Ratings apply for zero or negative gate voltage. Devices should not be tested for blocking capability in a manner such that the voltage supplied exceeds the rated blocking voltage. *Indicates JEDEC Registered Data

ELECTRICAL CHARACTERISTICS ($T_C = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Typ	Max	Unit
Peak Forward Blocking Voltage ($T_J = 125^\circ\text{C}$)	V_{DRM}	50	—	—	Volts
		100	—	—	
		200	—	—	
		400	—	—	
		800	—	—	
Peak Forward Blocking Current (Rated V_{DRM} @ $T_J = 125^\circ\text{C}$)	I_{DRM}	—	—	2.0	mA
Peak Reverse Blocking Current (Rated V_{DRM} @ $T_J = 125^\circ\text{C}$)	I_{RRM}	—	—	2.0	mA
Forward "On" Voltage ($I_{TM} = 24$ A Peak)	V_{TM}	—	1.7	2.2	Volts
Gate Trigger Current (Continuous dc) (Anode Voltage = 12 Vdc, $R_L = 100$ Ohms)	I_{GT}	—	5.0	30	mA
Gate Trigger Voltage (Continuous dc) (Anode Voltage = 12 Vdc, $R_L = 100$ Ohms)	V_{GT}	—	0.7	1.6	Volts
Gate Non-Trigger Voltage (Anode Voltage = Rated V_{DRM} , $R_L = 100$ Ohms, $T_J = 125^\circ\text{C}$)	V_{OD}	0.2	—	—	Volts
Holding Current (Anode Voltage = 12 Vdc)	I_H	—	6.0	40	mA
Turn On Time ($I_{TM} = 12$ A, $I_{GT} = 40$ mA dc)	t_{on}	—	1.0	2.0	μs
Turn Off Time (V_{DRM} = rated voltage) ($I_{TM} = 12$ A, $I_R = 12$ A)	t_{off}	—	15	—	μs
Forward Voltage Application Rate ($T_J = 125^\circ\text{C}$)	dv/dt	—	35	—	V/ μs

*Indicates JEDEC Registered Data

FIGURE 1 - AVERAGE CURRENT DERATING

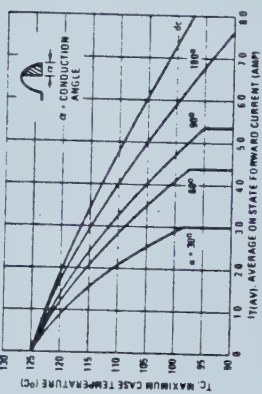
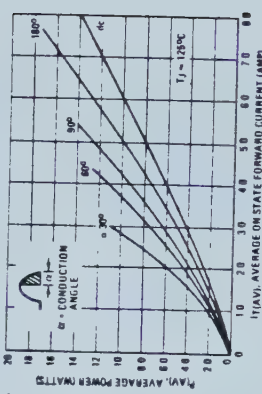


FIGURE 2 - MAXIMUM ON STATE POWER DISSIPATION



MJE350 (SILICON)

PLASTIC MEDIUM POWER PNP SILICON TRANSISTOR

- ... designed for use in line-operated audio and television applications and as low power, line-operated series pass and switching regulators.
- High Collector-Emitter Sustaining Voltage - $V_{CE(sat)} = 300 \text{ Vdc}$ @ $I_C = 1.0 \text{ mAdc}$
- Excellent DC Current Gain - $h_{FE} = 30-240$ @ $I_C = 50 \text{ mAdc}$
- Plastic Thermoplastic Package

0.5 AMPERE POWER TRANSISTOR PNP SILICON

300 VOLT
20 WATT



MAXIMUM RATINGS

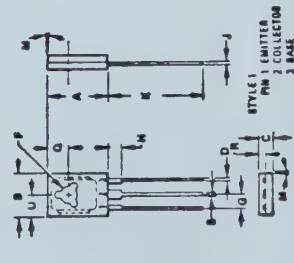
Rating	Symbol	Value	Unit
Collector-Emitter Voltage	V_{CE}	300	Vdc
Emitter Base Voltage	V_{EB}	3.0	Vdc
Collector Current - Continuous	I_C	500	mAdc
Total Device Dissipation @ $T_C = 25^\circ\text{C}$	P_D	20	Watts
Operating and Storage Junction Temperature Range	T_J, T_{stg}	-65 to +150	$^\circ\text{C}$

THERMAL CHARACTERISTICS

Characteristic	Symbol	Value	Unit
Thermal Resistance, Junction to Case	θ_{JC}	0.26	$^\circ\text{C/W}$

ELECTRICAL CHARACTERISTICS ($T_C = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Max	Unit
DC CHARACTERISTICS				
Collector-Emitter Sustaining Voltage ($I_C = 1.0 \text{ mAdc}, I_E = 0$)	$V_{CE(sat)}$	300	-	Vdc
Collector Cutoff Current ($V_{CE} = 300 \text{ Vdc}, I_E = 0$)	I_{CBO}	-	100	μA
Emitter Cutoff Current ($V_{EB} = 3.0 \text{ Vdc}, I_C = 0$)	I_{EBO}	-	100	μA
ON CHARACTERISTICS				
DC Current Gain ($I_C = 50 \text{ mAdc}, V_{CE} = 10 \text{ Vdc}$)	h_{FE}	30	240	-



DIM	MIN	MAX	MIN	MAX
A	10.80	11.00	0.425	0.433
B	7.45	7.75	0.293	0.309
C	2.41	2.60	0.095	0.103
D	0.75	0.75	0.030	0.030
E	0.75	0.75	0.030	0.030
F	0.75	0.75	0.030	0.030
G	0.75	0.75	0.030	0.030
H	0.75	0.75	0.030	0.030
I	0.75	0.75	0.030	0.030
J	0.75	0.75	0.030	0.030
K	0.75	0.75	0.030	0.030
L	0.75	0.75	0.030	0.030
M	0.75	0.75	0.030	0.030
N	0.75	0.75	0.030	0.030
O	0.75	0.75	0.030	0.030
P	0.75	0.75	0.030	0.030
Q	0.75	0.75	0.030	0.030
R	0.75	0.75	0.030	0.030
S	0.75	0.75	0.030	0.030
T	0.75	0.75	0.030	0.030
U	0.75	0.75	0.030	0.030
V	0.75	0.75	0.030	0.030
W	0.75	0.75	0.030	0.030
X	0.75	0.75	0.030	0.030
Y	0.75	0.75	0.030	0.030
Z	0.75	0.75	0.030	0.030

CASE 77-03

MJE350 (continued)

FIGURE 1 - DC CURRENT GAIN

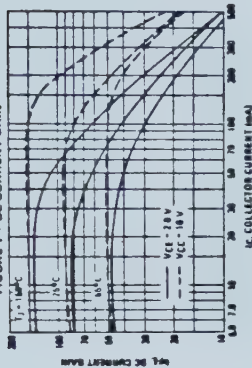


FIGURE 2 - "ON" VOLTAGES

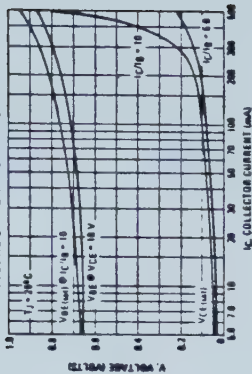


FIGURE 3 - ACTIVE REGION SAFE OPERATING AREA

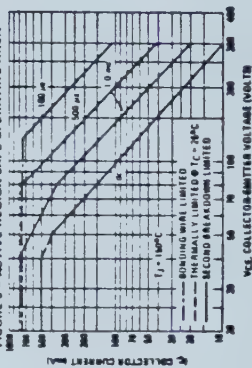


FIGURE 4 - TEMPERATURE COEFFICIENTS

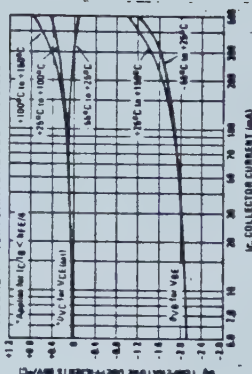
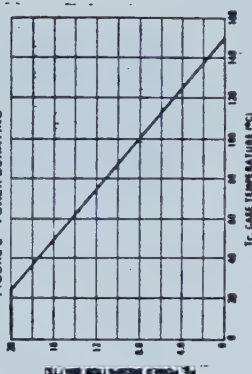


FIGURE 5 - POWER DERATING



There are two limitations on the power handling ability of a transistor: average junction temperature and second breakdown. Safe operating area curves indicate $I_C - V_{CE}$ limits of the transistor that must be observed for reliable operation. I.e., the transistor must not be subjected to greater dissipation than the curves indicate. The base of Figure 3 is based on $T_{J(max)} = 150^\circ\text{C}$. T_C is variable depending on conditions. Exceeding these limits will cause thermal stress, thermal limitations will reduce the power that can be handled to values less than the limitations imposed by second breakdown. (See AN-418).

MPS-U07 (continued)

ELECTRICAL CHARACTERISTICS (T_A = 25°C unless otherwise noted)

OFF CHARACTERISTICS				
Characteristic	Symbol	Min	Type	Max
Collector-Emitter Breakdown Voltage (1)				
(I _C = 1.0 mA; I _B = 0)	BV _{CEO}	100	—	—
Emitter Base Breakdown Voltage				
(I _E = 100 μA; I _C = 0)	BV _{EB0}	4.0	—	—
Collector Cutoff Current				
(V _{CE} = 80 Vdc; I _E = 0)	I _{CBO}	—	—	100 nA
ON CHARACTERISTICS				
DC Current Gain (1)				
(I _C = 50 mA; V _{CE} = 1.0 Vdc)	h _{FE}	80	110	—
(I _C = 250 mA; V _{CE} = 1.0 Vdc)		30	65	—
(I _C = 500 mA; V _{CE} = 1.0 Vdc)		—	33	—
Collector-Emitter Saturation Voltage (1)				
(I _C = 250 mA; I _B = 10 mA)	V _{CE(sat)}	—	0.18	0.4
(I _C = 250 mA; I _B = 25 mA)		—	0.1	—
Base-Emitter On Voltage (1)				
(I _C = 250 mA; V _{CE} = 5.0 Vdc)	V _{BE(on)}	—	0.78	1.2
SMALL-SIGNAL CHARACTERISTICS				
Current Gain - Bandwidth Product (1)				
(I _C = 200 mA; V _{CE} = 5.0 Vdc; f = 100 MHz)	f _T	80	175	—
Output Capacitance				
(V _{CE} = 10 Vdc; I _E = 0; f = 100 kHz)	C _{ob}	—	6.8	12

(1) Pulse Test: Pulse Width ≤ 300 μs, Duty Cycle ≤ 2.0%.

FIGURE 1 - DC CURRENT GAIN

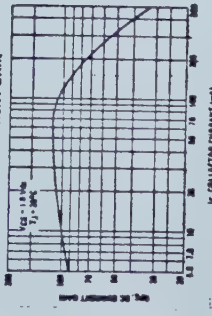


FIGURE 3 - DC SAFE OPERATING AREA

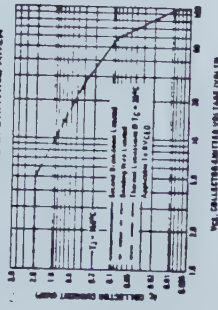


FIGURE 2 - "ON" VOLTAGES

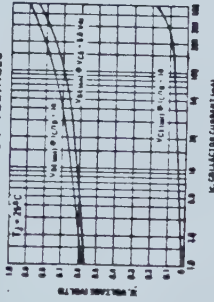
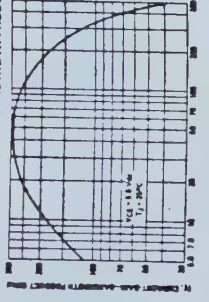


FIGURE 4 - CURRENT GAIN-BANDWIDTH PRODUCT



The data of Figure 3 is based on T_{junction} = 150°C. T_C is variable depending on conditions. At high case temperatures, thermal resistance will reduce the power that can be handled to values less than the limitations imposed by ambient breakdown.

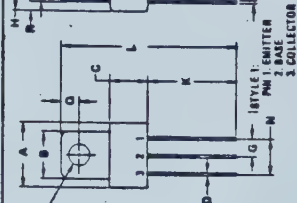
There are two limitations on the power handling ability of a transistor. Junction temperature and second breakdown. Safe operating area curves indicate I_C - V_{CE} limits of the transistor that must be observed for reliable operation. I.e., the transistor must not be subjected to greater dissipation than the curves indicate.

MPS-U10 (SILICON)

NPN SILICON ANNULAR TRANSISTOR

... designed for high-voltage video and luminance output stages in TV receivers.

- High Collector-Emitter Breakdown Voltage - BV_{CEO} = 300 Vdc (Min) @ I_C = 1.0 mA
- Low Collector-Emitter Saturation Voltage - V_{CE(sat)} = 0.75 Vdc (Max) @ I_C = 30 mA
- Low Collector Base Capacitance - C_{ob} = 3.0 pF (Max) @ V_{CB} = 20 Vdc



DIM	MIL	MM	INCHES
A	0.14	0.32	0.012
B	0.04	0.04	0.001
C	0.04	0.04	0.001
D	0.04	0.04	0.001
E	0.04	0.04	0.001
F	0.04	0.04	0.001
G	0.04	0.04	0.001
H	0.04	0.04	0.001
I	0.04	0.04	0.001
J	0.04	0.04	0.001
K	0.04	0.04	0.001
L	0.04	0.04	0.001
M	0.04	0.04	0.001
N	0.04	0.04	0.001
O	0.04	0.04	0.001
P	0.04	0.04	0.001
Q	0.04	0.04	0.001
R	0.04	0.04	0.001
S	0.04	0.04	0.001
T	0.04	0.04	0.001
U	0.04	0.04	0.001
V	0.04	0.04	0.001
W	0.04	0.04	0.001
X	0.04	0.04	0.001
Y	0.04	0.04	0.001
Z	0.04	0.04	0.001

CASE 182-02

MPS-U10 (continued)

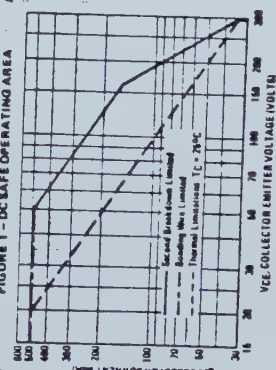
MPS-U10 (continued)

ELECTRICAL CHARACTERISTICS ($T_A = 25^\circ\text{C}$ unless otherwise noted)

Characteristic	Symbol	Min	Max	Unit
OFF CHARACTERISTICS				
Collector-Emitter Breakdown Voltage (1)	BV_{CEO}	300	—	Vdc
Collector Base Breakdown Voltage	BV_{CBO}	300	—	Vdc
Emitter Base Breakdown Voltage	BV_{EBO}	8.0	—	Vdc
Collector Cutoff Current	I_{CB0}	—	0.2	μA
Emitter Cutoff Current	I_{EB0}	—	0.1	μA
ON CHARACTERISTICS				
DC Current Gain				
$I_C = 1.0 \text{ mAdc}$; $V_{CE} = 10 \text{ Vdc}$	h_{FE}	26	—	—
$I_C = 10 \text{ mAdc}$; $V_{CE} = 10 \text{ Vdc}$		40	—	—
$I_C = 30 \text{ mAdc}$; $V_{CE} = 10 \text{ Vdc}$		40	—	—
Collector-Emitter Saturation Voltage	$V_{CE(sat)}$	—	0.76	Vdc
Base-Emitter On Voltage	$V_{BE(on)}$	—	0.86	Vdc
DYNAMIC CHARACTERISTICS				
Current Gain - Bandwidth Product (1)	f_T	60	—	MHz
Collector-Base Capacitance	C_{cb}	—	3.0	pF

(1) Pulse Test: Pulse Width $\leq 300 \mu\text{s}$, Duty Cycle $\leq 2\%$.

FIGURE 1 - DC SAFE OPERATING AREA



The Safe Operating Area Curve indicates I_C - V_{CE} limits below which the device will not enter second breakdown. Collector load lines for specific circuit must fall within the applicable Safe Operating Area to avoid causing a catastrophic failure. To insure operation below the minimum T_J , power temperature derating must be observed for both steady state and pulse power conditions.

FIGURE 2 - DC CURRENT GAIN

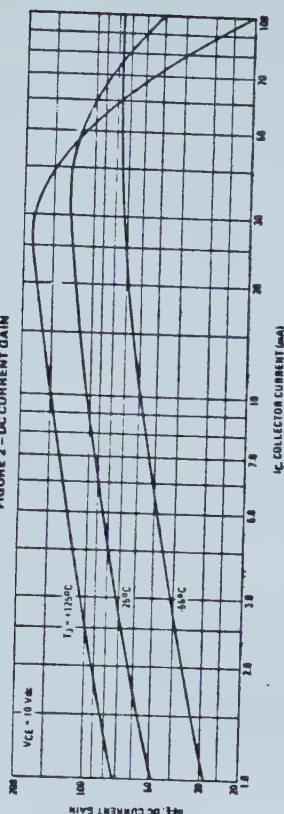


FIGURE 3 - CAPACITANCES

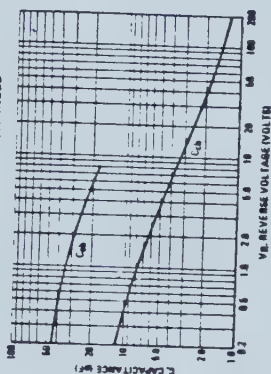


FIGURE 4 - CURRENT GAIN-BANDWIDTH PRODUCT

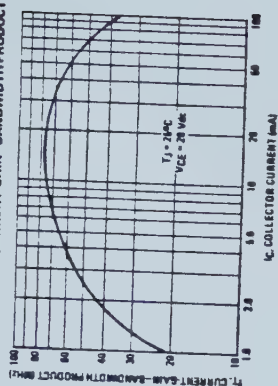
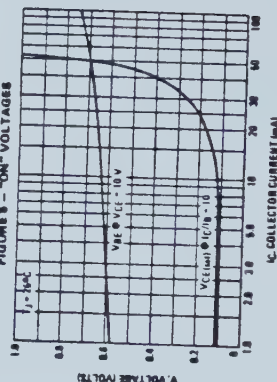


FIGURE 5 - "ON" VOLTAGES



8.2 Control Program Source Code

```

/*file: check.c
/*function:  checkcurrent
author:      chris
date:        27.10.81
version:     2.0
description:  reads and writes the a/d converter line
               and sets a flas ,if the current limit
               was exceeded before returning the value
               of the conversion.
*/

#include "exdefs.c"

checkcurrent()
{
    int conv;
    extern currflas;
    extern float volt;
    conv = a2d(0);
    if (conv >=LIMIT){
        currflas = 1;
    }
    volt = (float) a2d(1)*2/10;
    return(conv);
}

```



```

/*file:      ov.c
function:    overcurrent
author:      chris
version:     2.0
date:       3.11.91
description:  this routine writes the last extinguishing word
              then prints out the error message and terminates.
*/

#include "extdefs.c"

overcurrent/extins:
int extins;
{
    DR11.1->dr_buf = extins ;
    printf("Overcurrent-condition occurred\nProgram terminated\n");
    DR11.1->dr_buf = RESET;
    exit(-1);
}

```



```

/*file:      compare.c
function:    compare
author:      modified from C-manual p125
version:     1.4
date:        4.11.81
description: compares the characterstrings with a standard set of keywords
              and returns an integer that gives the position in that set.
routines
called:      strcmp(library routine)
*/

static char *commtab[] = {"catalos","display","help","new frequency",
                          "parameters","quit","reverse","run","softstart","stop","torque"};

compare(s,n)
char *s;
int n;
{
    int low,mid,high,cond;

    low = 0;
    high = n - 1;
#ifdef DEBUG
    printf("compare: n %d high %d\n",n,high);
#endif

    while(low <= high){
        mid = (low+high)/2;
#ifdef DEBUG
        printf("mid %d\n",mid);
#endif
        if ((cond = strcmp(s,commtab[mid],3)) < 0)
            high = mid - 1;
        else if (cond > 0)
            low = mid + 1;
        else
            return(mid);
    }
    return(-1);
}

```



```

/*file:      convert.c
function:    convert
author:      chris
date:        18.11.81
version:     5.5
description:  takes the input new frequency
              and converts it to the appropriate cycletime
              then computes the numbers for the timings
              intervals alpha and beta.

functions called:  none
*/
convert(freal)
int freal;
{
    extern int *beta,*alpha;
    int corr;
    static int x,z;
    double xcr,bf;
    a = 30000;
    b = 1000; /* a and b are the constants that determine */
              /* the slope of the V/f characteristic */
              /* for frequencies < 30 Hz */
    corr = 250; /* correction for system-dependent delay */
#ifdef DEBUG
    printf("convert\007 : \n");
    printf("give numbers for a/b \n");
    while(getch('%f %f',&a,&b) !=2)
        ;
#endif
    if(freal < 9)
        freal = 9;
    x = 1000000/(6 * freal);
    if(freal < 30){
        z = 2500 +(int) (a/freal) - b;
        y = (int)x - z - corr;
    }
    if(30 <= freal){
        if(freal < 60 ){
            z = 2500; /*conduction interval is constant for f <= 60 Hz */
            y = (int)x - z - corr; /*pause between extins-pulse and next firing pulse */
        }
        else{
            y = 10; /* minimum pause, Caution when trying to change !!! */
            z = (int)x - corr;
        }
    }
    beta = &z;
    alpha = &y;
#ifdef DEBUG
    /***/ printf ("%alpha=%d\n", *alpha) ;
    /***/ printf ("LEAVE CONVERT\n") ;
#endif
    return;
}

```



```

/*file: exdefs.c
external definitions for the main program  */

#include <stdio.h>
#include <vector.h>
#include <dt2762.h>
#include <dt2749.h>
#include <drv11.h>

#define COM20 046 /* output word for +A-B+C (0010 0110) */
#define SCOOTER 045 /* +A-B-C (0010 0101) */
#define CORK 051 /* +A+B-C (0010 1001) */
#define FOZZY 031 /* -A+B-C (0001 1001) */
#define KFP*IT 032 /* -A+B+C (0001 1010) */
#define FUGGY 024 /* -A-B+C (0001 0110) */
#define E60N 0200 /* word for extinguishing + row (1000 0000) */
#define FRIDA 0100 /* - row (0100 0000) */

#define LIMIT 2000 /* limit for dc-current */
#define RTC_LM1 0110 /* set clock rate to 1Mhz/mode 0 */
#define RTC_LM0 010 /* same rate as above but no interrupt */
#define RTC_LG0 1 /* start bit for clock */
#define RESET 0 /* RESET- constant */
#define VLDONE 0150 /* clock interrupt vector */
#define CLYOVER 120 /* clock-overflow */
#define MASK 0600 /* clock overflow mask */

#define N*F*E 11

```



```

/*file:      interrupt.c
function:    interrupt
author:      chris
date:        24.11.91
version:     1.0
description:  makes sure that thyristors are off if external
              interrupts occur then prints message and terminates
              program
*/

#include "e:defs.c"
interrupt()
{
    int i;
    /* print %int_word%

    IF11_1->dr_xbuf = %int_word%;
    for(i=0;i < 2;i++)
        ;
    IF11_1->dr_xbuf = RESET;
    printf("Routine convert: floatingpoint error; program terminated \n");
    exit(-1);
    */
}

```



```

/*file:          main.c
function:        main
author:          chris schmidt
version:         3.6
date:           19.11.81
description:     main routine
routines called: convert,start,compare
*/

#include "exdefs.c"
float curr,volt;
int *beta,*alpha; /* pointers to conduction,nonconduction-intervals */
int freq,ref,startflas,tflas,forward;
float k_i,k_p,k_d;

main(fp)
FILE *fp;
{
    int c;convert();
    char comm[20];
    curr = 0;
    freq = 60;
    volt = 0;
    forward = 1;
    k_p = 0.1;
    k_i = 0.9;
    k_d = .0015;

    convert(freq);
    printf("\007Hi there!\n Welcome to the motor-control program.\n \
For instruction type 'catalog'\n");
    for(;;){
        while(scanf("%s",comm) != EOF){
            if((c = compare(comm,NKEYS)) != -1){
                switch (c) {
                    case 0 :
                    case 2 : /*catalog or help */
                        printf("You only have to type the first three \
characters of each command \n \
catalog - produces \
this output \n display - shows values of voltage,frequency and current \n \
help - does the same as catalog \n new frequency - read in new frequency \n \
run - starts motor \n soft - perform softstart \n torque - starts with \
selectable torque profile \n \
parameters - reads in new parameters for the PID-controller \n \
reverse - reverses the motor spin \n \
quit - stops execution \n stop - does the same \n");
                        break;
                    case 1: /* show the values */
                        printf("The current values are:\n DC-voltage: %.3s\
frequency: %3d motor current: %s \n K-P: %s K-I: %s \
K-D %s \n",volt,freq,curr,k_p,k_i,k_d);
                        break;
                }
            }
        }
    }
}

```



```

case 3:          /* new frequency */
    printf("Input new frequency settings.\n Allowable \
range : 10 - 120 (inteseer values)\n ");
    if(scanf("%3d",&frea) == 1){
        if(frea > 9){
            convert(frea);
            start(fr);
        }
    }
    else
        printf("newfrequency: bad input\n ");
    break;
case 4 :          /* gain factors for control */
    printf("Gain factors K_P, K_I, K_D (in float-format) ??\n");
    if(scanf("%f %f %f",&k_p,&k_i,&k_d) != 3)
        printf("wrong input\n");
    break;
case 5 :          /*stop or quit */
case 9 :
    printf("bye\n ");
    exit(0);
case 6 :          /* reverse direction */
    forward = !forward;
    start(fr);
    break;
case 7 :          /*start the machine */
    start(fr);
    break;
case 8 :          /* softstart */
    printf("Give maximum frequency. Allowable range 10-120 Hz \n ");
    if(scanf("%3d",&iref) == 1){
        startflas = 1;
        frea = 10;
        convert(frea);
        start(fr);
    }
    else
        printf("start: error \n ");
    break;
case 10 :         /*torque profile */
    iflas = 1;
    frea = 10;
    convert(frea);
    start(fr);
    iflas = 0;
    break;
}
}
else
    printf("\007\007error: invalid command, try again \n");
printf("Command..\n ");
}
}

```



```

/*file:      speed.c
function:    speed
author:      chris
date:        21.11.81
version:     2.4
description:  reads rotor frequency from a/d channel 2
              and determines if any corrections to the feedins
              frequency should be made.If there was a change
              it sets the flas and returns.

routines
called:      none
*/

#include 'exdefc.c'

speed()
{
extern int freq,ref,channel;
extern float k1,k2,k3,k4;
float rotfreq;
float t_sample,prop,intesrel,deriv,delta_out;
static float delta_in[3];

t_sample = freq;
t_sample = 1/t_sample;
rotfreq = 22d/2/6;
if(rotfreq <= freq)
    delta_in[2] = rotfreq + 0.05 * rotfreq;
else
    delta_in[2] = rotfreq - 0.05 * rotfreq;
prop = k1 * (delta_in[1] - delta_in[2]);
intesrel = k2 * t_sample * (ref - delta_in[2]);
deriv = k3 * (2*delta_in[1] - delta_in[2] - delta_in[0])/t_sample;
delta_out = prop + intesrel + deriv;
if(delta_out > 0){
    if(delta_out > 0.2 * rotfreq) /*check if sliplimit exceeded */
        delta_out = 0.1 * rotfreq;
}
else if (delta_out < 0){ /* check for nesative sliplimit */
    if(delta_out < 0.2 * (-rotfreq))
        delta_out = 0.1 * (-rotfreq);
}
freq = freq + (int)delta_out;
delta_in[0] = delta_in[1];
delta_in[1] = delta_in[2];
#ifdef DEBUG
printf("speed : P %f I %f D %f del %f frequency %d\n",prop,intesrel,deriv,delta_out,freq);
#endif
}

```



```

/*
File:      start.c
function:   start
author:     chris
version:    5.5
date:       1.11.81
description: selects the words to be written out by the pulse-function
              and in the intervals between pulses performs a current-check
              ;determines if there was input from the keyboard, and does
              the speed control.

routines
called:     check;overcurrent;convert;pulse.
*/

#include "endsfs.c"

static int word1[] = {
    EGON,FRIDA,EGON,FRIDA,EGON,FRIDA};
static int word2[] = {
    GONZO,SCOOTER,COOK,FOZZY,KERMIT,PIGGY};
static int word3[] = {COOK,SCOOTER,GONZO,PIGGY,KERMIT,FOZZY};
int currfles=chancefles;
int *int1word;

start(%s
FILE *fp)
{
    register extern fire;
    int w1z=1;
    int count,*wp1,*wp2;
    int interrupt();
    extern float curri;
    extern int startflag,tfiles,ireq,ref,forward;
    extern int *alpha, *beta;
    struct vector *vector;

    wp1 = word1;
    if(forward)
        wp2 = word2;
    else
        wp2 = word3;    /*for reversed firing order*/

```



```

chanseflas = currflas = 0;
vector->v_ps = PL0;
vector->v_pc = V_FPE;
count = 0;
z = 0;
DR11_1->dr_xbuf = FRIDA;      /*charge commutation capacitor */
attach(vector->v_pc,interrupt,vector->v_ps);
#ifdef DEBUG
                                /*this debussing statement allows */
printf ("START\007\007!\n") ;
printf("now what ??");        /*to check, if the capacitor is */
while(scanf("%d",&z) != 1)      /*charged properly,          */
    ;                          /* To continue, Just type a number */
#endif
for(;;){
    for (i=0;i<=5;i++){
        extins = *(wr1+i);
        fire = *(wr2+i);
        DR11_1->dr_xbuf = fire;
        DT2769->rt_xbuf = -*beta; /*conduction time */
        DT2769->rt_losr.w = RTCLRO RTCLGO;
        J=0;
        while(J<3)
            J++;
        DR11_1->dr_xbuf = RESET;
        intLword = extins;

        if(i==0){
            x = checkcurrent();
            curr = (float) x/200;
            if(currflas)
                overcurrent(extins);
        }
        if(i == 1){
            if (empty(fp) == 0){ /* something has been written on the terminal */
                DR11_1->dr_xbuf = extins;
                for(J = 0;J< 2;J++) /* wait for 100us */
                    ;
                DR11_1->dr_xbuf = RESET;
                return;          /* back to main routine */
            }
        }
    }
}

```


B30336